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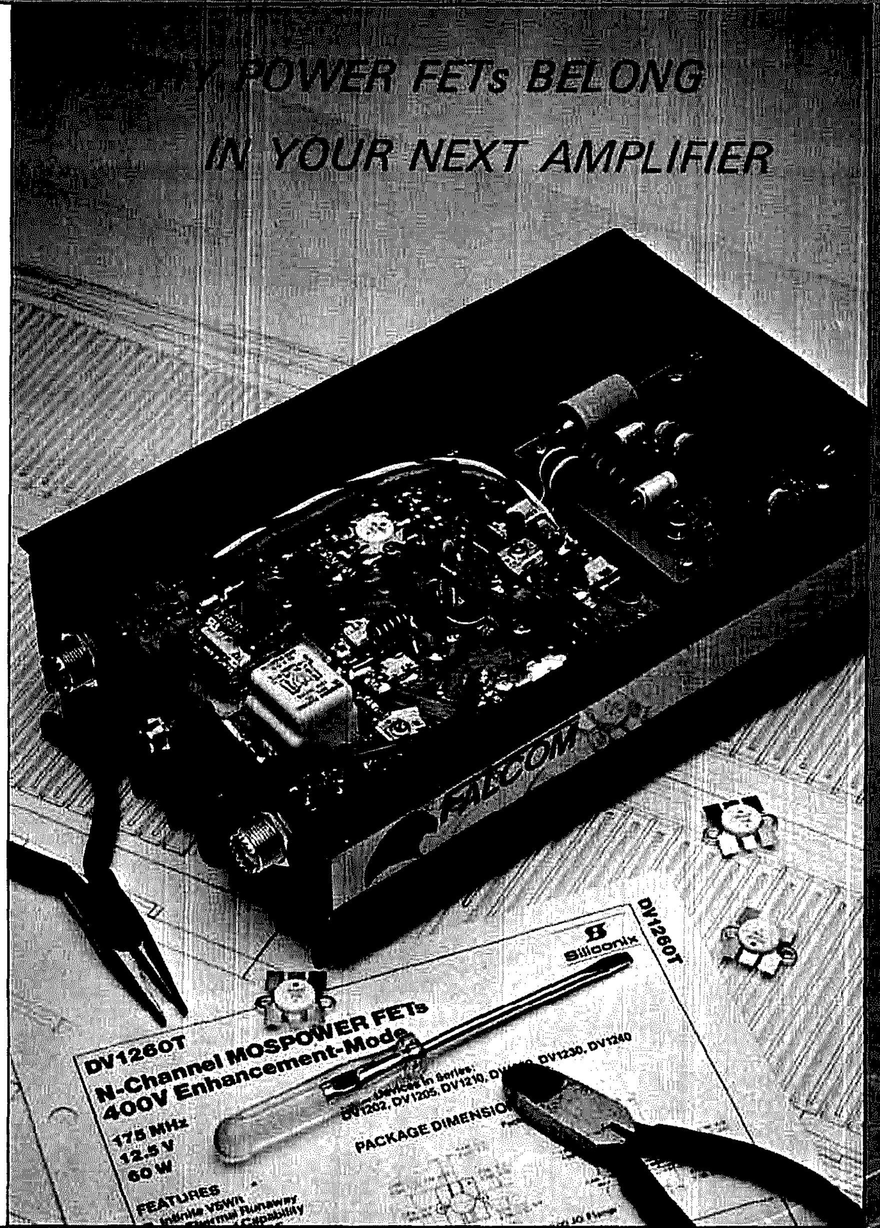

hr

focus on communications technology

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JANUARY 1984

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contents

COVER: Siliconix, Inc.

Lost Weekend

With winter rapidly approaching, this past weekend was supposed to be spent cutting, splitting, and stacking wood for our stoves; operating for a few hours in the contest (CQ WW CW); and compiling the results of the last few questions of the September reader survey. Who could predict that my high tree-supported wire antennas would be belted with almost nonstop gusty winds and twice need repair — and a five-band trap vertical need a midnight pruning? I didn't count the number of times I ran back and forth from the shack to the vertical, but it numbered in the *scores*. (There wasn't much difference in temperature between the bitter cold outside and the shack . . . I hadn't yet cranked up the stove.)

After relaxing from the contest, I looked forward to leisurely compiling the thoughtful responses to the question that asked readers to tell us how they thought *ham radio* might be improved.

I read each and every answer to that question at least twice. Many comments were just what we'd expected. While some were indeed eye-opening, none were shocking. Here's what I found:

About your Amateur interest. The average reader of *ham radio* holds at least an Advanced class license (Extra, 33 percent; Advanced, 39 percent; General, 16.5 percent; Technician, 7 percent; Novice, 1.4 percent; no license, 1.4 percent) and has been licensed for 19 years. (33 readers in our sample have each been licensed for over 51 years.)

You operate mostly on SSB/AM. (CW and FM modes are just about tied for second place.) More than half the time you spend on 80 through 10, though the VHF bands attract the second largest group. A little over half of you build your own equipment; considering the level of interest indicated in construction articles, probably more of you would like to. Of those who build, half build from kits and half from "scratch."

How does this translate into hours spent on all facets of Amateur Radio? According to our poll, half our readers spend at least ten hours per week in operation or related activities.

Over half of you own personal computers. And of the half that don't, a third plan to purchase one within the coming year. What isn't clear is how many of you who own computers use them in Amateur Radio applications. (*Let us know.*)

About *ham radio*. What do readers want from us? Specifically, you asked for more articles on antennas (the #1 favorite subject overall), receivers, and using computers for Amateur Radio. There was a general request — almost across the board — for more construction articles, and almost as much interest in articles of a more theoretical nature. Many different subjects attracted your attention (not surprising in view of your diversified backgrounds, occupations, and interests); we'll use this information to plan future issues.

To which magazines do your subscribe? *ham radio*, of course (98.5 percent, and to a lesser extent, the three other brand names. (This comes as no surprise, considering that it was *our* readership that was polled.)

Which best suits your needs? This question really brought out the diversity of our readership. No, you don't think that *ham radio* is tops in terms of reporting on station activities, contesting, politics, news stories, or nostalgia. You buy the other books for that. That's fine. We at *ham radio* aren't trying to be a something-for-everybody magazine. We're just trying to provide the best technical ham magazine you can obtain. That has always been our charter.

I want to thank all who responded, and especially the hundreds of readers who took the time to expand their views in additional comments and even lengthy letters. (One reader even sent an hour-long audio tape.) The quantity and quality of positive suggestions supplied would be highly gratifying to any editor; as the months and years unfold, I'll do my best to put your advice to work. One major reader request has already been incorporated: starting in this issue, Joe Reisert, W1JR, will contribute a monthly column on VHF/UHF. Joe's many technical and operating accomplishments should be of interest to old and new readers alike.

As this issue went to press, we heard of the untimely passing of Vic Clark, W4KFC, President of the ARRL, over the Thanksgiving weekend. To me, Vic personified Amateur Radio. He was a dedicated ham, a kind person who was always thinking of ways to improve the hobby. He will be missed.

Rich Rosen, K2RR
Editor-in-Chief

presstop de W9JUV

ARRL PRESIDENT VIC CLARK, W4KFC, PASSED AWAY SUDDENLY November 25. He had complained of chest pains Thanksgiving day, and went to the hospital for examination. He seemed to be doing OK, but was still in the hospital for observation Friday night when a sudden heart attack claimed him. Vic had suffered some health problems in the past few years.

Vic, A "Ham's Ham," Served Amateur Radio With Distinction for most of his life. He had been an ARRL director, played key roles on FCC/Amateur advisory committees and in various International Amateur Radio Union activities, yet always found time to get on the air. A top CW contest operator and DXer, W4KFC's call was for many years invariably found among the top scores in the Sweepstakes and DX frays. Just a week before his death Vic was made a Fellow of the Radio Club of America, "for leadership in Amateur Radio organization, including WARC preparation and implementation." When a League director recently expressed concern that Vic was pushing himself too hard he replied that he wouldn't feel he'd done his part if he didn't die "in the saddle."

Memorial Services Were Held November 30 in Washington, DC; contributions in Vic's memory can be made to the ARRL Foundation c/o ARRL. He's survived by his wife, Hester, WA4PAE, and six children, three of them also Amateurs. W4KFC, a Silent Key at 66.

New ARRL President Is Carl Smith, WØBWJ, as automatically provided by the League By-Laws. Succeeding Carl as First Vice President is Larry Price, W4RA. A retired airline captain who learned aerial gunnery from Lt. Barry Goldwater back in 1940, Carl is well known and respected in both domestic and international Amateur circles. His work on the 1979 WARC also led to constructive working relationships with key people at the FCC, as well. No radical changes in ARRL direction are expected under Carl's direction.

ARRL'S WISH FOR VEC COMPENSATION WAS ANSWERED when the House and Senate passed the FCC Authorization (funding) Bill just before Thanksgiving. Smooth political manuevering by Senator Barry Goldwater, who'd reversed his earlier opposition to any fees in the program, resulted in the fee-permitting amendment being attached to the FCC bill. It was then passed without dissent. FCC Chairman Fowler also supported the pro-fee change.

ARRL Can Now Proceed To Develop Its Own VEC Role; its Executive Committee had put its program on hold pending outcome of the fee question. However, it could still be some time before fees are actually incorporated into the rules. With such delicate questions as fee allocation and acceptable accounting procedures unanswered, the Commission will probably decide that a formal rule-making procedure—usually a matter of months—is called for.

In The Meantime A Number Of Other Organizations including educational institutions have expressed their interest in becoming VECs. It now appears a distinct possibility that, at least in some areas, we could end up with more than one VEC!

10-METER REPEATERS WILL REMAIN LIMITED TO 29.5-29.7 MHZ, at least for the immediate future. Acting on PR Docket 83-485, the Commission decided the interference potential with satellite downlinks and other 10 meter users was too great to justify any change. The ARRL, which initially supported expansion, had filed against any change at this time pending results of the further Notice of Proposed Rule Making on phone band expansion.

Phone Band Expansion Has Been Pushed To The Back Burner, with current FCC resources occupied with the volunteer exam program, Amateur involvement in rules enforcement, and of course the "No-Code" license. Latest Washington readings indicate action on the no-code license is very close, possibly as a "Christmas present" to the Amateur community.

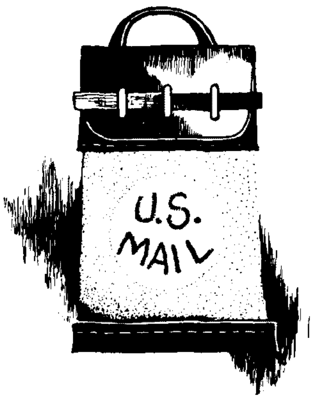
W5LFL DID GET ON THE AIR FROM THE SPACE SHUTTLE, with a full-quieting signal into a hand-held and rubber duckie. At press time the first confirmed QSO was WA1JXN/7 in Montana on orbit 40, though he may also have worked the West Coast on an earlier pass. Lots of media exposure has also been reported. QSL and SWL cards should go c/o ARRL with an SASE.

PRESSURE ON THE 220-MHZ BAND IS STILL INCREASING, with the FCC's Office of Science and Technology now suggesting that the 220-225 MHz allocation be the subject of an FCC Notice of Proposed Rule Making. Though Amateurs are currently the sole users, the band is actually shared with government and land mobile on a co-prime basis.

Band Usage Could Be The Principal Determinant of its future, both in quantity and quality. One way that might help preserve the band would be to make it Amateur Radio's prime high tech "workplace," with, for example, packet radio and various wideband techniques. This is one of the approaches the ARRL is planning to take.

WB6JAC'S CONVICTION FOR TRANSMITTING OBSCENITY on the Amateur bands has been reversed in the U.S. Ninth District Court of Appeals. Though the court did not dispute that Burton had transmitted "obscene" language, it said the government prosecutors had failed to show his actions aroused any "prurient interests!" His convictions for operating without a license still stand. However, such cases may be easier to sustain in the future. Another amendment to the FCC Authorization Bill extends Commission authority over "Dial-A-Porn" telephone businesses since their content is potentially available to children, and that's justification for not providing it First Amendment protection. Since Amateur Radio is operated primarily in homes or family automobiles, the same sanctions could be applied.

Rick Cooper, The Former Exponent Of Unlicensed "HF" Operation, has surfaced again as a result of the Burton reversal. In a rambling letter to the Amateur Radio media, he and his "Communications Attorney Service" promised to support Burton in a lawsuit against "...ARRL, FCC and all radio hams who conspired to deprive Mr. Burton of his constitutional rights...."



comments

10-meter beacon

Dear HR:

The 10-meter beacon described in the September, 1983 issue of *ham radio* ("Comments," page 13) has been moved from Niantic, Connecticut, to just outside of Rochester, New York. It is now about 10 miles (16 km) south of Rochester at 43° 02′ N, 77° 41′ W, in grid square FN 13 using the Maidenhead Grid Locator system. The power is still 4 watts output and the antenna is a dipole up about 20 feet (6 meters). The beacon, on 28.286 MHz CW, is on the air 24 hours a day.

W. Keith Hibbert, KA1YE
Rush, New York

TOM remembered

Dear HR:

I read your "Reflections" column in October, 1983, *ham radio*. I don't think I've ever been so pleased to read *anything* in Amateur Radio publications.

I was fortunate. I read "TOM" early in my exposure to Amateur Radio. Whenever a bad practice would make an appearance, TOM would be there with the way to get back on the track. It seemed that all Amateurs had tremendous respect for TOM, and as a result, he had a tremendous effect upon the behavior of the majority of licensed operators.

As I remember, I was perhaps one of the first to complain about lists. I wrote to — was his name Newkirk? — the DX editor of *QST* at the time and registered my protest against what I perceived to be a cancer growing on DXCC. Of course, nothing happened except that as time passed, DXCC became less of an accomplishment. Naturally, if the real competition is removed from DXCC (or any other award), the award becomes less desirable to the true DXer. I wonder what sense of accomplishment one receives from Honor Roll status when he knows that the certificates should be made out in the name of the do-good MC?

I am proud of my DXCC, 5B DXCC, WAZ, etc. I have 308 confirmed and every last one is self-earned! To disprove the equipment dodge, I have only modest equipment, but it has been put together with DX in mind, using a tri-band beam. But I expect that if one wants to work DX, he must equip himself with DX gear just as an automobile racer equips himself. (I don't ever remember TOM saying it was wrong to run legal power — or to use gain antennas.)

With the MC type of DX, you stand in line and take your turn! This should never have been allowed to happen. I blame all of the DX commentators for forgetting that DX is a competition and that the chase is where the fun is. I'm very pleased that *ham radio* has brought the issue out of the closet. Perhaps continued exposure of this practice will make it dry up and disappear. If it is ridiculed sufficiently — and often — I feel sure that many Amateurs will avoid it, and in so doing regain the thrill of DXing.

Walter Camuso, W1ESN
Vero Beach, Florida

avoiding splatter

Dear HR:

W5XW's excellent letter (December, 1983) commenting on the poor waveform of the sidetone of some keyers is correct. A poor waveform would cause splatter if fed into the audio input of a transmitter. My own homebrewed circuit uses a Twin-T audio oscillator and produces a good waveform output that has been checked on an oscilloscope. And my on-the-air tests have shown that there was far less interference caused than when you yell AHHHHHH into the microphone, or the side bands caused by normal conversation. I did check on a buddy's commercial keyer's sidetone, and it was poor, just as Bob said.

The relay technique I described was not used in the audio input circuit, but was used in the keying circuit because the frequency of the relay was far too low for a good audio signal. It acted just like the dots of my old Vibroplex, when I tested the idea years ago. And as I said, I did prefer the automatic keyer because it gave a steadier output, as would be expected.

The low-duty cycle technique, used with a dummy load, is very effective when tuning up because before you tune for your plate current dip, your plate dissipation can temporarily be quite high, and at times reach dangerous levels. Another good reason for first using a proper dummy load is that when your final amplifier is tuned up, you don't have to touch its controls when you switch to your antenna tuner. This technique naturally eliminates the interaction problem between transmitter and tuner controls that is so often present when you tune up directly without first using a dummy load. The more you use your dummy load, the less QRM you make on the bands.

So to avoid splatter as mentioned by W5XW, do as he suggests, unless your audio side tone output is a good sine wave; just use dots and key your transmitter directly. (And thanks, Bob, for your precautionary letter. I had not been aware of the poor quality of sidetone wave shapes of the kind of units he discussed until he brought it to my attention.)

William Vissers, K4KI
Cocoa Beach, Florida

Note: As this issue went to press, Siliconix announced the sale of the RF portion of its MOSFET power transistor business to the PHI Division of M/A-Com, in Torrance, California. Siliconix will continue to supply these parts until M/A-Com is in full production of the product. — Editor

power FETs:
trend for VHF amplifiers

Use these MOSFETS for better thermal stability, lower noise, easier matching, and higher voltage operation

The dull black fins of heatsinks have all but replaced the glow of vacuum tubes as the distinguishing visual feature of power amplifiers designed for Amateur use. Bipolar junction power transistors offer many advantages and have, among amplifier designers, become the device of choice over vacuum tubes in many applications. However, just as bipolar junction transistors have largely replaced vacuum tubes as technology has advanced, it now appears that bipolar transistors are being challenged by field effect transistors (specifically RF power MOSFETs) in the RF power amplifier field.

In this article we will discuss some of the features that make the RF power MOSFET (variously known as VMOS, TMOS, DMOS, etc. by different manufacturers) an impressive RF power amplifier. We will also touch on some of the problems that accompany their use. Measurements made on several actual VHF mobile power amplifiers will serve to illustrate the discussion. The amplifiers examined are 50 and 100 watt 2-meter units and a 100 watt unit for 220 MHz.

advantages

When compared to conventional bipolar transistors, the RF power MOSFET offers the following advantages:

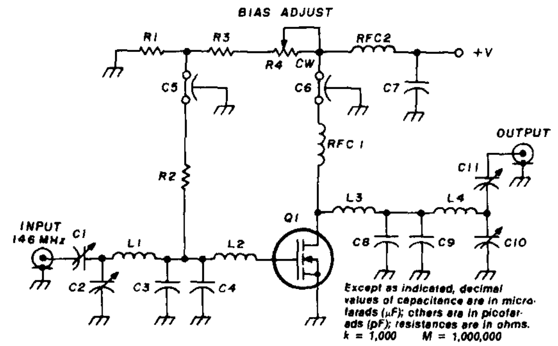
Thermal stability. The current gain (Beta) of a typical bipolar transistor increases with temperature. As a result, the collector current can increase with temperature, which results in still higher temperatures and even higher currents, and can lead to thermal runaway. Careful circuit design is required to prevent this problem. On the other hand, at high power levels the transconductance of a power FET decreases with temperature, and an increase in temperature results in a decrease in current, which tends to be self-stabilizing. This stabilization also applies across the chip and serves to prevent destructive phenomena characteristic of a bipolar device: current hogging, hot spotting, and secondary breakdown. Also, no internal source ballasting resistors — with their inherent gain reduction, increased parasitic capacitance, and increased fabrication costs — are required.

Another advantage of this temperature characteristic is the ability to parallel several power FETs without the need for careful device matching. With VHF amplifiers, the number of devices that can be paralleled is limited more by the physical problem of additional parasitic reactances than anything else.

Low noise. A power FET generates far less broadband noise, typically 10 dB better, than a comparable bipolar transistor. This is due partially to the absence of a forward-biased junction and its associated shot noise. This low noise has the potential for significant reduction of transmitter noise levels, which can be beneficial at repeater sites or at any location where other equipment is operating in close proximity.

Low spurious. The transfer characteristic of a typical power FET displays no abrupt changes in shape. This means that when biased for Class AB operation, as in a typical linear power amplifier application, there will be lower high-order inter-

By Daniel Peters, ex DL4VJ, Falcon Communications, Suite 400-550, 2995 Woodside Road, Woodside, California 94062, and **Robert S. Larkin, W7PUA**, Janel Laboratories, 33890 Eastgate Circle, Corvallis, Oregon 97330



parts list

designation	*description*
C1	4-40 pF mica trimmer
C2	2-20 pF mica trimmer
C3,C4	100 pF leadless mica
C5,C6	1000 pF feedthrough
C7	1000 pF disc ceramic
C8,C9	100 pF leadless mica
C10	2-20 pF mica trimmer
C11	4-40 pF mica trimmer
L1	0.025 μH
L2	0.003 μH
L3	0.003 μH
L4	0.025 μH
Q1	Siliconix DV1260T
R1	1 K, ¼ watt
R2	10 K, ¼ watt
R3	2.2 K, ¼ watt
R4	10 K, trimmer pot
RFC1,2	9 turns No. 20 wire, 5/32 inch diameter

fig. 1. 2-meter test amplifier schematic.

modulation products than in a similarly operated bipolar transistor.

Input impedance. The power FET gate is essentially an MOS capacitor. At low frequencies this results in much higher input impedances than the equivalent bipolar device. At VHF the ratio is less favorable. However, the FET device still tends to look capacitive, as opposed to the inductive reactance presented by bipolar devices, which simplifies the design of the input matching networks.

Reduced feedback. Power FETs have reduced internal feedback paths. The higher input impedance results in gate drive voltages several times as high as typical base drive voltages, with two benefits: first, the voltage induced across the source impedance affects the input voltage of the FET proportionally less than the equivalent voltage across the emitter inductance in a bipolar transistor circuit. Second, the effect of reverse transfer capacitance, already low in the power FET, is further reduced by the lower voltage gain.

Parameter changes. The transfer parameters of the RF power FET are quite insensitive to power level. This translates into smooth tuning of the input and output, along with continuous input/output curves, and contrasts with bipolar devices that tend to require retuning at each power level. Bipolar devices also tend to have jumps in power output due to parameters changing with power level.

Simplified circuit design. Gate leakage current is in the nanoampere or sub-nanoampere range resulting in essentially no bias power being used. Thus, simple, low-power bias circuits can be utilized. In addition, the negative temperature coefficient of the FET allows the use of bias supplies without the complex temperature compensation schemes common to bipolar designs. In some higher power designs, it is still desirable to use temperature compensation. However, the compensation is to reduce variations in circuit performance with temperature, not to protect the devices.

Higher operating voltage. Although the practical circuits we will consider in this article are oriented toward mobile VHF power amplifiers, the newer high voltage FETs reaching the market present exciting possibilities for base station use. As supply voltage increases, current decreases and impedance levels increase. At a fixed power level, doubling the voltage halves the current and quadruples the impedance. Increased impedance reduces the effect of parasitic inductance elements and makes the internal leads of the transistor a less critical part of the matching networks. Capacitor values in the matching networks become more reasonable and bypassing gets much easier. Finally, increased impedance allows easier broadband transformer design.

Ruggedness. The previously discussed thermal properties of the FET are often mentioned as the reason for their toughness when compared to bipolar devices.

Another equally important factor contributing to this quality is the FETs' voltage ratings. For example, the gate in the Siliconix devices, specified for 13.6 volts service, will withstand at least 30 volts with respect to the source or drain, and the drain will withstand at least 45 volts with respect to the source. Devices with even higher voltage ratings are available.

The result of these high breakdown voltages and favorable temperature characteristics is an amplifier that can withstand considerable abuse. Neither reasonable amounts of excessive input power nor high VSWR loads will cause any problems for the FET.

commercially available RF power MOSFETS

The preceeding discussion of the advantages of RF power MOSFETs is relevant only if you can pur-

chase the devices and actually use them in an amplifier. **Table 1** lists some important parameters of a sample of the RF power MOSFETs available from Siliconix, Motorola, and Acrian, respectively. (Other manufacturers offer RF power MOSFETs, so your choice is not limited to these three. Also note that these are only partial lists, and that the selection continues to increase as time passes.)

The purpose of **table 1** is simply to illustrate the ranges of some of the devices available and to list the performance of the particular devices under certain sets of circumstances. The devices are useful under conditions other than those listed. For example, the Motorola devices specified at 28 volts perform very well, although at reduced gain and power levels, at 13.6 volts. The higher power levels listed are PEP ratings. Thermal considerations make steady-state power outputs at higher power levels impractical. Higher power levels require the paralleling of devices, fans, and other parts.

building a test amplifer

The Siliconix DV1260T, described in **table 1**, is a particularly interesting device for use in VHF mobile power amplifiers. The rest of this article will be devoted to examining the use of this device in several practical circuits.

The first step was to build a lab model of a single-transistor 2-meter amplifier in order to verifiy the Siliconix data and explore possible problems in the areas of stability, DC voltages, gain, and RF matching. The design goal was a 50-watt amplifier usable for both FM and SSB.

Fig. 1, a schematic of the amplifier, reveals a straightforward approach. Both input and output matching networks use double L sections. The double L sections result in lower losses than single section networks. The input network transforms a gate impedance of 1 ohm to 50 ohms. The 1-ohm gate impedance is essentially resistive at these frequencies because the input capacity is near series resonance with the lead inductance. If done in a single L sec-

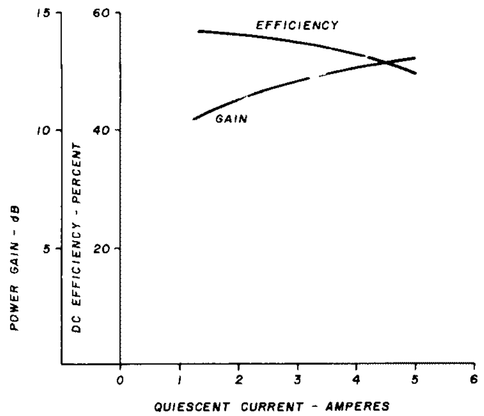


fig. 2. Gain and efficiency of 2-meter test amplifier as function of quiescent current at 50-watt output level.

table 1. Selected RF power MOSFETs available from Siliconix, Motorola, and Acrian.

device	price	typical performance at 2 meters
	Siliconix RF power MOSFETs	
DV1220S	$21.90	20 watts out with 10 dB gain at 13.6 volts
DV1240U	31.03	40 watts out with 9 dB gain at 13.6 volts
DV1260T	44.80	60 watts out with 9 dB gain at 13.6 volts
DV2820S	20.75	20 watts out with 12 dB gain at 28.0 volts
DV2840S	44.20	40 watts out with 12 dB gain at 28.0 volts
DV2880U	84.35	80 watts out with 10 dB gain at 28.0 volts
DV28120T	100.80	120 watts out with 10 dB gain at 28.0 volts
DVD150T	100.80	150 watts out with 12 dB gain at 120.0 volts
	Motorola power MOSFETs	
MRF136	16.00	20 watts out with 15 dB gain at 28 volts
MRF171	35.00	50 watts out with 15 dB gain at 28 volts
MRF172	65.00	80 watts out with 12 dB gain at 28 volts
MRF174	88.00	120 watts out with 11 dB gain at 28 volts
MRF150	92.00	150 watts out with 10 dB gain at 50 volts
	Acrian power ISOFETs	
VMIL20FT	33.00	20 watts out with 13 dB gain at 28 volts
VMIL40FT	45.00	40 watts out with 13 dB gain at 28 volts
VMIL60FT	65.00	60 watts out with 13 dB gain at 28 volts
VMIL80FT	77.00	80 watts out with 13 dB gain at 28 volts
VMIL120FT	105.00	120 watts out with 10 dB gain at 28 volts

tion, a loaded Q of about 10 is required. This produces a 1 dB loss for coils of reasonable Q. The double L section allows loaded Qs of about 3 and the two sections produce a loss of about 0.5 dB. Incidentally, because the input and output impedance transformations were very similar at the power level of interest, similar components were used for both networks.

Testing of the amplifier was done at supply voltages of both 13.6 and 16 volts. 16 volts was included for potential base station use because the performance of the DV1260T improved a bit at higher voltages. The amplifier was tested at 146 MHz.

An exploration of the effects of quiescent current was made. If the quiescent current is too high, the device will overheat from static dissipation. If the quiescent current is too low, gain and linearity suffer. **Fig. 2** shows the gain and efficiency of the amplifier as a function of quiescent current. A quiescent current of 3 amperes was used for most of the remainder of the tests.

Fig. 3 shows the gain of the amplifier as a function of the output level. At 16 volts a gain of over 15 dB is achieved at low power levels, dropping to about 10 dB at 75 watts output. At 50 watts the gain was measured at 12.2 dB. The lower curve shows the expected decrease in gain at 13.6 volts.

Fig. 4 demonstrates the effect of output power level on efficiency. At first glance it might appear that the efficiency is greater at 13.6 volts than at 16 volts. However, we must look at the efficiencies at equivalent points with respect to maximum power output. For example, the amplifier operating at 16 volts and 75 watts has about the same gain and linearity as when operating at 13.6 volts and 50 watts and as might be expected, the efficiency at 16 volts and 75 watts is greater than for 13.6 volts and 50 watts.

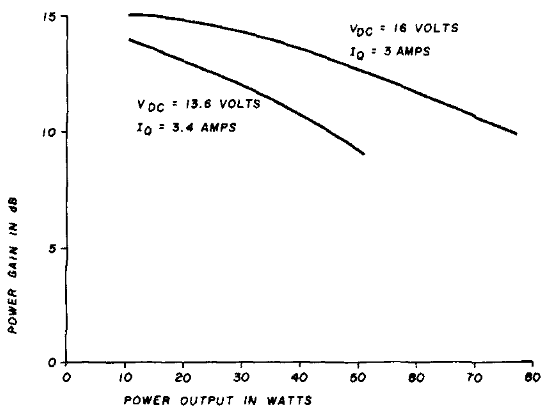


fig. 3. Gain of amplifier as function of output level.

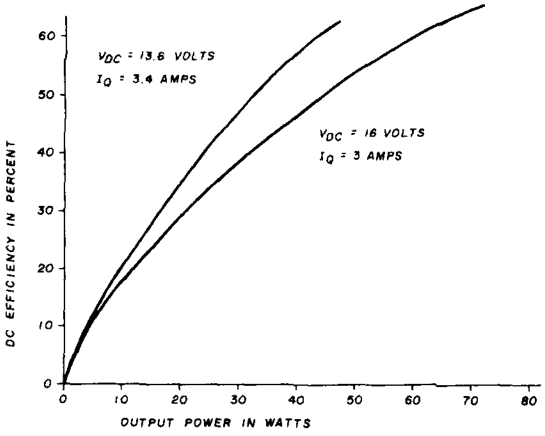


fig. 4. DC efficiency as a function of output power level.

Measured at 25 watts, the test amplifier showed an input VSWR of just under 2. This was with the tuning adjusted for maximum output. This value would seem adequate for working with most Amateur transceivers.

Other tests were run and the following basic conclusions were reached. The DV1260T device performed essentially as claimed by Siliconix. A single device is quite capable of 75 watts when operating at 16 volts and 50 watts when operating at 13.6 volts. Looking forward, a pair of these devices would appear to be suitable for 100 to 150 watts when operating from 16 volts. When operating from 13.6 volts, a capability of 100 watts would seem practical.

a practical amplifier

The next step was to take the test amplifier and turn it into a practical amplifier by adding T/R switching and control circuitry. An amplifier of this power level will be used primarily to boost the power level of the ubiquitous two-meter handheld transceiver (HT) when used in an automobile, so we added a few other features: an adjustable regulated power supply to power the HT and save on batteries, an audio amplifier to boost the HT's sound level, and provision for plugging in a receiver preamplifier. **Fig. 5**, a schematic of the finished unit, represents a complete handheld transceiver (HT) accessory package. Because this is not intended to be a construction article, the schematics are supplied for functional guidance only; in the interest of brevity we will describe the circuit in terms of features rather than in specific detail. We hope it will serve as a source of ideas to those of you who will design their own FET amplifiers.

The unit was built on two PC boards, an RF board,

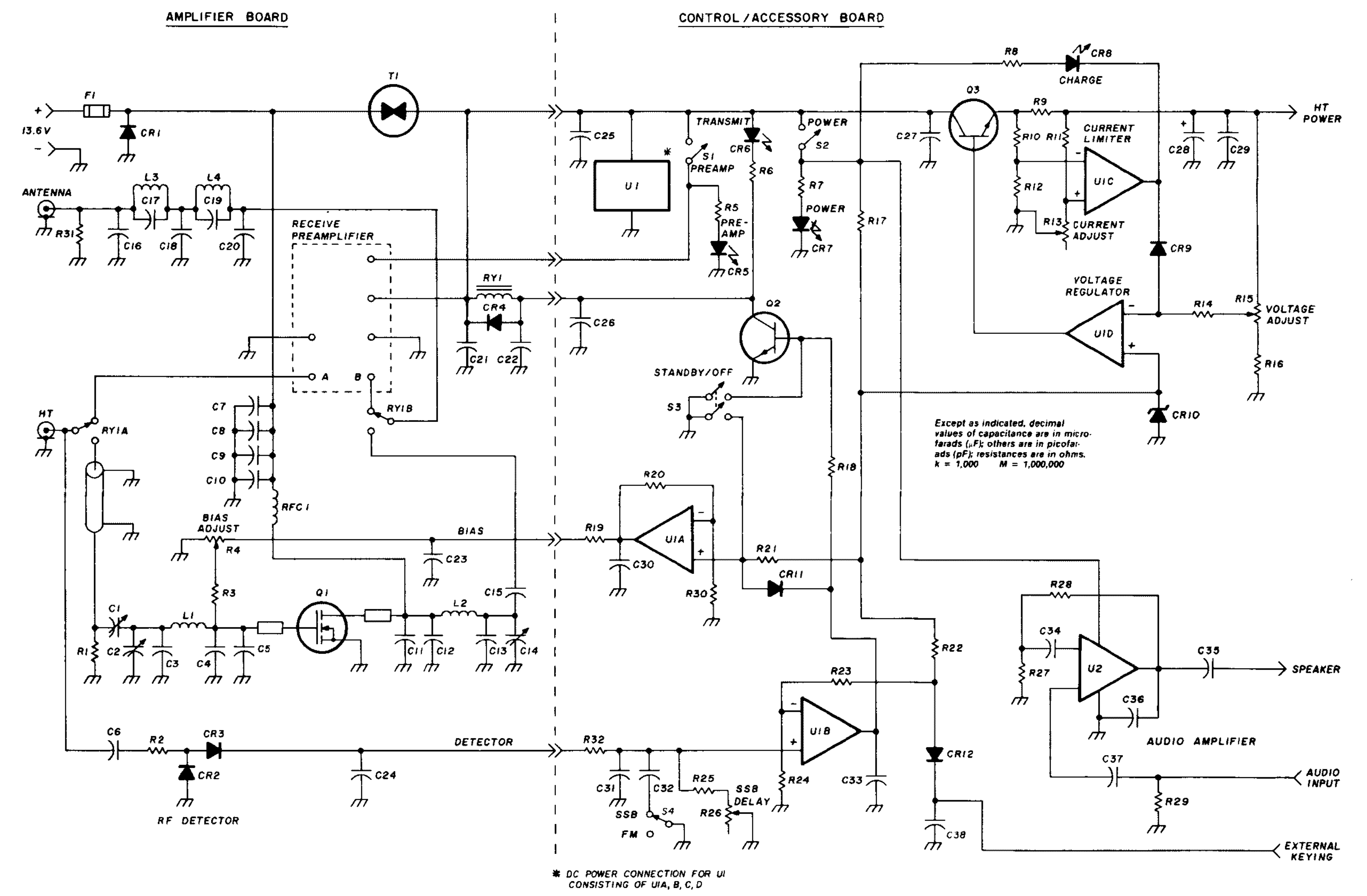


fig. 5. 2-meter HT amplifier and accessory unit schematic.

parts list for 2-meter amplifier/ accessory package

designation	*description*
C1,C2	2-20 pF mica trimmer
C3	10 pF disc ceramic NPO
C4,C5	100 pF Unelco
C6	1.5 pF disc ceramic NPO
C7,C8	470 pF disc ceramic
C9	1 µF tantalum
C10	1000 µF 16 volt
C11	100 pF Unelco
C12	22 pF mica
C13	10 pF disc ceramic NPO
C14	2-20 pF mica trimmer
C15	15 pF mica
C16,C17	10 pF disc ceramic NPO
C18	15 pF mica
C19,C20	10 pF disc ceramic NPO
C21,C22, C23,C24,C25	470 pF disc ceramic
C23,C24,C25,C26 C29,C30,C31,C33	470 pF disc ceramic
C27,C28	10 µF/16 volts
C32	22 µF/25 volts
C34	100 µF/16 volts
C35	100 µF/16 volts
C36	0.22 µF
C37	4.7 µF/16 volts
C36	470 pF disc ceramic
CR1	MR750 diode
CR2,CR3	1N4148 diode
CR4	1N4001 diode
CR5,CR6, CR7,CR8	LED indicator
CR9	1N4148 diode
CR10	5.6 volt Zener
CR11,CR12	1N4148 diode
F1	10 amp fuse
L1	3 turns No. 20 wire 5/32 inch diameter
L2	2 turns No. 20 wire 5/32 inch diameter
L3,L4	2½ turns No. 20 wire 5/32 inch diameter
R1	470 ohm, ¼ watt
R2	100 ohm, ¼ watt
R3	10 K, ¼ watt
R4	10 K, trimmer
R5	560 ohm, ¼ watt
R6	470 ohm, ¼ watt
R7	560 ohm, ¼ watt
R8	270 ohm, ¼ watt
R9	0.33 ohm, 1 watt
R10,R11, R12,R13	100 K, ¼ watt
R14	10 K, ¼ watt
R15	10 K trimmer
R16	10 K ¼ watt
R17	560 ohm, ¼ watt
R18	4.7 K, ¼ watt
R19	560 ohm, ¼ watt
R20	82 K, ¼ watt
R21	100 K, ¼ watt
R22	47 K, ¼ watt
R23,R24,R25	10 K, ¼ watt
R26	100 K trimmer
R27	2.7 ohm, ¼ watt
R28	270 ohm, ¼ watt
R29	22 ohm, ¼ watt
R30	10 K, ¼ watt
R31	100 K, ¼ watt
R32	560 ohm, ¼ watt
RFC1	9 turns No. 20 wire 5/32 inch diameter
S1,S2	SPST push-push switch
S3	DPST push-push switch
S4	SPST push-push switch
T1	thermostat, 175° F open
U1	LM324
U2	LM383
RY1	DC DPDT relay

and a control/accessory board. The RF board contains the amplifier and matching networks, a TR relay, a lowpass filter, and a receiver preamplifier. The preamplifier, designed especially for this project by Janel Laboratories, supplies 10 dB of gain for use with those HTs whose receivers need a little bit "extra". It is a plug-in unit and is replaced by a jumper across points A and B when the preamplifier is not needed.

The matching networks perform the same impedance transformations as on the test amplifier. The networks are cascaded L and T sections. Multiple sections are used for the same reasons as explained in the discussion of the test amplifier, above. The inductor for the L sections next to the transistor consists of the transistor leads and a short length of microstripline on the PC board.

The five-pole output lowpass filter has a cut-off frequency of about 200 MHz. Two filter response "zeros" are introduced at about 292 MHz as second harmonic suppressors. The zeros are the result of capacitors C17 and C19. The second harmonic is more than 65 dB below the fundamental. Higher harmonics are further down and not measurable with the equipment used.

Fig. 6 is the PC board pattern for the front and back of the RF board and **fig. 7** is the parts layout.

Fig. 8 shows power output versus power input. It is a little bit less than the output of the test amplifier because of the losses in the relay and lowpass filter.

The amplifier performs smoothly. Its input/output characteristics exhibit none of the discontinuities or hysteresis common to bipolar power transistors. The tuning is very smooth and the tuning characteristics do not vary much with power level, as with bipolar transistors. Varying tuning, power level, and load VSWR while monitoring spectral output revealed no trace of spurious outputs.

The amplifier is well behaved and fully stable in any operating enviroment. It is possible to create oscillations by reactively terminating the input with no load on the output. This condition does not occur in normal operation.

A thermostat that opens at 175 degrees Fahrenheit is mounted on the heatsink near the amplifier transistor. Although you should always use more than adequate heatsinks to prevent overheating, there is always the possibility that even a well-protected unit might be covered by something — a carelessly thrown sweater, for example — that could cut off air flow. In such an event, in which overheating would be possible, the thermostat could prevent expensive damage.

The control/accessory board supplies the necessary switching and control circuitry for the amplifier and accessory circuits to increase the utility of the HT in the mobile environment.

An RF detector on the amplifier board supplies a voltage when the HT is keyed. This is sensed by U1B and turns on Q2 which controls the TR relay. U1B also controls U1A, which supplies a regulated voltage for biasing the amplifier. Without bias the amplifier draws no current. A regulated supply keeps the amplifier specifications more constant as the supply voltage varies.

When S4 is in the "FM" position, the application of bias and the closing of the TR relay is essentially instantaneous, following the sensing of an RF signal. When the HT ceases transmitting, the bias is removed and the TR relay opens, also essentially instantaneously. When S4 is in the "SSB" position, C32 is added to the circuit and while it doesn't slow the turn-on significantly, it does slow the turn-off time. This delay, determined by C32 and, R25 and R26, is

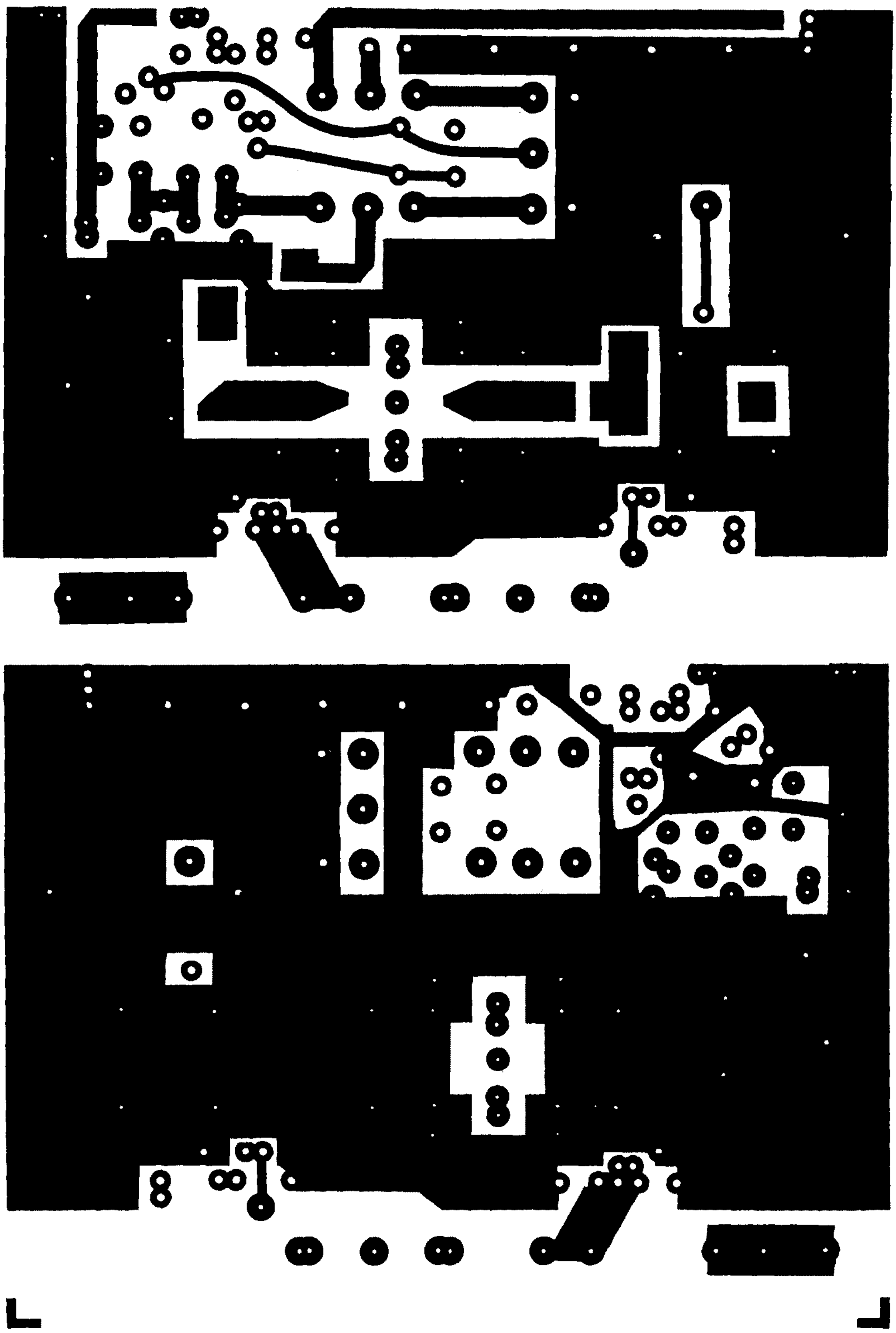


fig. 6. PC board patterns for 50-watt 2-meter amplifier: front (above), back (below).

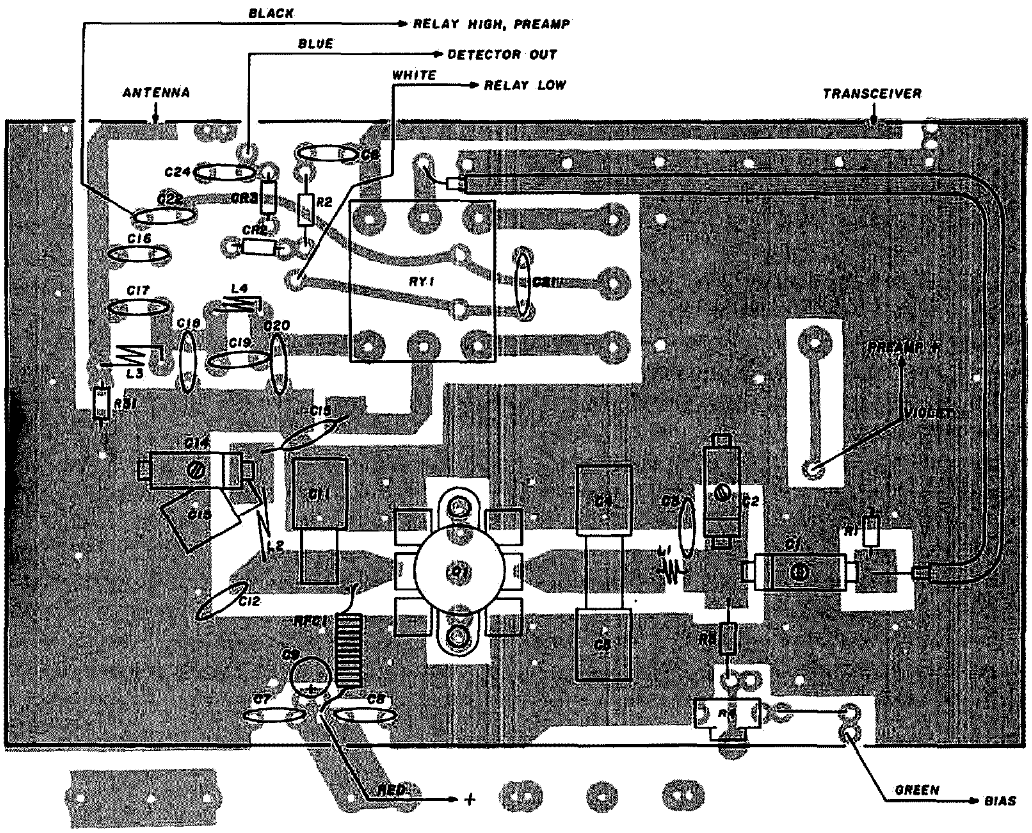


fig. 7. Component layout for 50-watt 2-meter amplifier.

added so that the unit doesn't turn off between words because of the lack of a carrier on SSB.

Most multi-mode amplifiers on the market have this switch. Since the amplifiers are biased for linear operation in both FM and SSB modes, these switches have nothing to do with the linearity of the amplifier and serve only to insert a delay in one RF sensing circuit in the SSB mode to prevent the amplifier from keying in and out between words. The delay is adjustable and a compromise setting must be found. If you set the delay time long enough so that the amplifier doesn't cut in between words, it may take an uncomfortably long time to switch to receive after you let up on the PTT switch.

RF switching is convenient but a better solution is to wire the amplifier for direct keying. An external keying line is supplied on this and most other amplifiers for this purpose.

U1C, U1D, and Q3 form a current-limited voltage regulator to power the HT. This circuit can be used in two different ways. In one, it will power the HT instead of the HT batteries. R15 is adjusted to set the output voltage at the rated voltage of the HT and R13

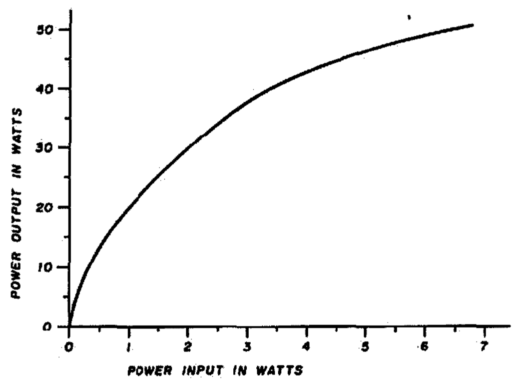


fig. 8. Output power versus input power for 2-meter HT amplifier.

is adjusted so that the output current limits at a little over the transmit current drain of the HT. The power supply can deliver a little over one ampere.

The alternative is to use the power supply as a charger for the HT batteries. For this, adjust R15 to

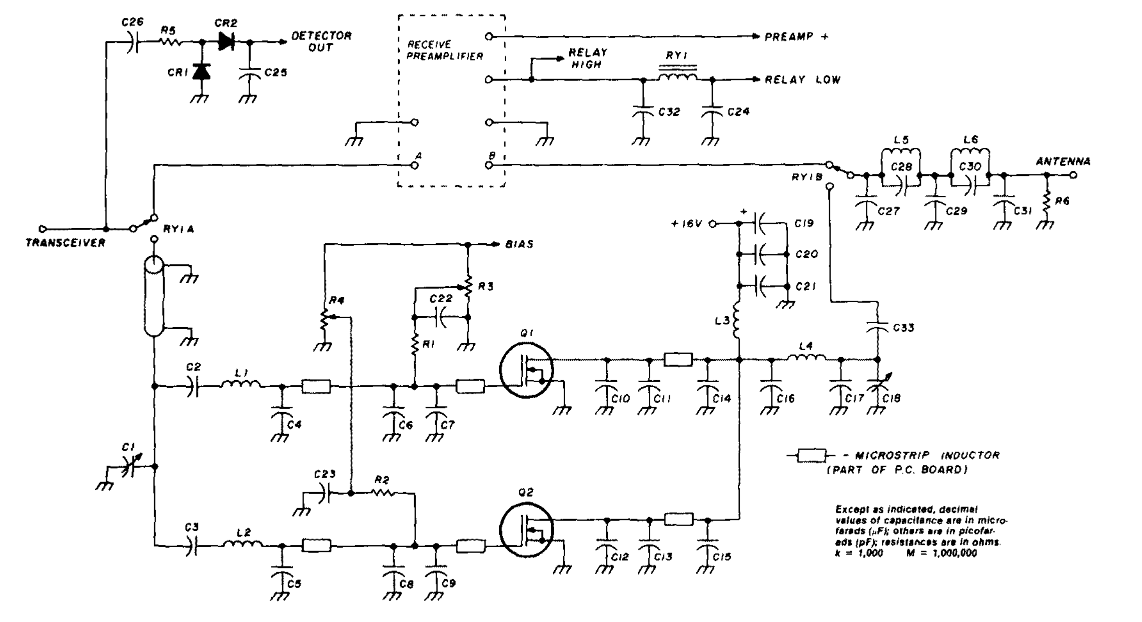


fig. 9. 100-watt 2-meter amplifier schematic.

set the output voltage to the fully charged voltage of the HT batteries and set R13 to the recommended charging current for the batteries. The batteries will be charged at this current until they reach full charge, at which time the power supply automatically switches to constant voltage and reduces the charge rate to whatever is needed to maintain full charge. CR8 will glow when the supply is in the constant current mode, confirming that the batteries are under charge. (See the January, 1983, issue of *ham radio* for an excellent discussion of this subject.)[1] In the first mode, discussed in the previous paragraph, CR8 indicates an overload.

The final major accessory circuit is the one containing U2. This is a speaker amplifier to boost a low HT audio output signal to a level capable of driving a speaker that can deal with the high background noise in a vehicle. The LM383 is capable of supplying sufficient current. However, it is limited to a peak voltage swing of a little less than 1/2 the supply voltage. So use a low impedance loudspeaker if you want to make a lot of noise. The LM383 will typically deliver 4.7 watts to a 4-ohm load with a 13.2 volt supply. It will deliver 7.2 watts to a 2-ohm load, but only 2.4 watts to an 8-ohm load.

Making the control/accessory board separate from the amplifier board provides some flexibility in mounting. When the amplifier is mounted within easy reach of the operator, the control/accessory board mounts in the same cabinet as the amplifier.

parts list for fig. 9

designation	description
C1	2-20 pF mica trimmer, Arco 402
C2,C3	22 pF dipped mica
C4,C5,C6,C7, C8,C9,C10, C11,C12,C13	100 pF Unelco J101 type
C14,C15,C16	47 pF Unelco J101 type
C17	24 pF Unelco J101 type
C18	6-60 pF mica trimmer, Arco 404
C19	1 μF 35-volt tantalum capacitor
C20,C21,C22 C23,C24,C25	470 pF disc ceramic
C26	1.5 pF disc ceramic, NPO
C27,C28	10 pF disc ceramic, NPO
C29	15 pF disc ceramic, NPO
C30,C31	10 pF disc ceramic, NPO
C32	470 pF disc ceramic
C33	560 pF dipped mica
CR1,CR2	1N4148
L1,L2	7 turns No. 20 enamelled wire 5/32-inch ID
L3	5 turns No. 16, 1/4-inch enamelled wire, 5/32-inch ID
L4	2 turns 1/8-inch copper wire, 3/16-inch ID
L5,L6	3 turns No. 20 enamelled wire, 5/32-inch ID
Q1,Q2	Siliconix DV1260T
R1,R2	10K 1/4 watt
R3,R4	10K linear pot
R5	100 ohm, 1/4 watt
R6	100K, 1/4 watt
RY1	DC DPDT relay

When the amplifier is mounted out of reach, such as in the trunk, the control/accessory board is removed from the amplifier chassis, mounted in a small remote control cabinet, and connected by cable.

higher power

The next step was to design a pair of amplifiers using 2 parallel DV1260Ts. The first amplifier was designed for 2 meters and the second for 1-1/4 meters.

As expected, the 2-meter amplifier was similar to the amplifiers just described. At a supply voltage of

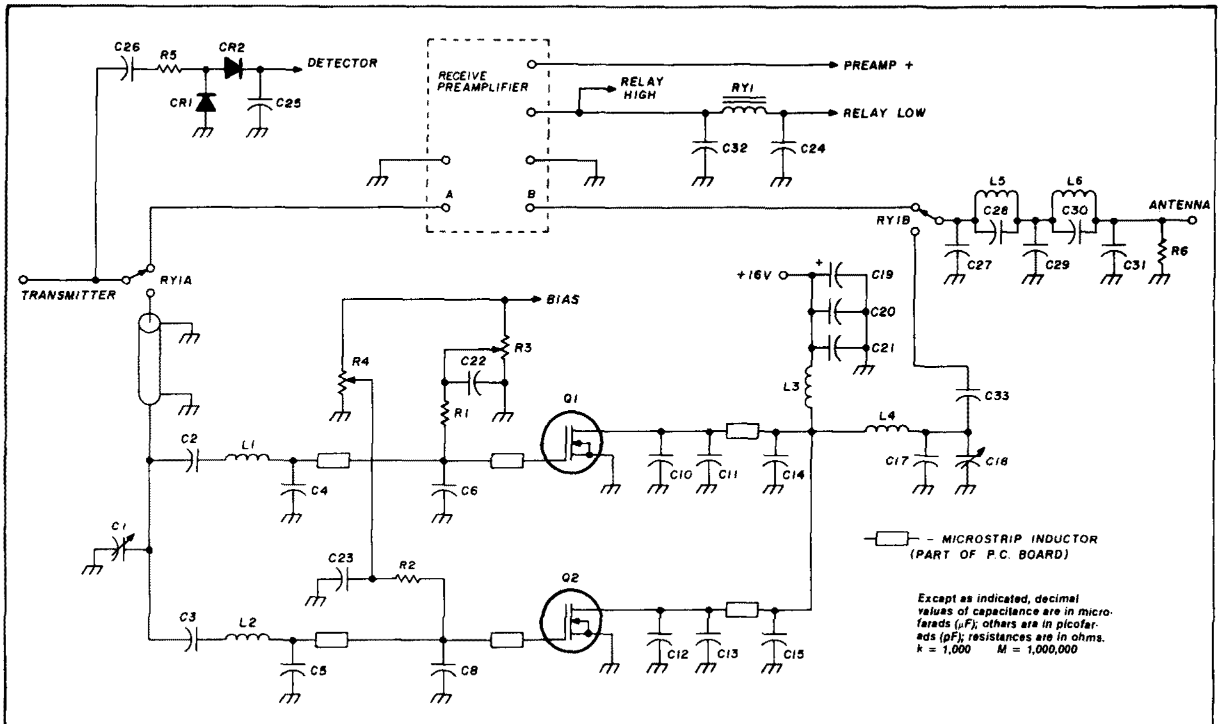


fig. 10. 100-watt 220-MHz amplifier schematic.

16 volts, 100 watts of output was obtained with a little less than 10 watts of input. The gain and efficiency were lower at 13.6 volts. Performance curves are not included because all you have to do is use the single transistor amplifier curves and adapt for the doubling of power.

A single stage amplifier of this gain and power output will most likely be used with a mobile transceiver. A catalog search revealed that most new transceivers have a 25-watt output, so a design which supplied 100 watts output with 25 watts of input was pursued. **Fig. 9** shows the amplifier schematic. Only the amplifier portion is shown because the control board is a simplified version of the one described for the single transistor amplifier, less the audio amplifier and the voltage regulator.

The amplifier consists of two single-device amplifiers connected in parallel. Each single-device amplifier is designed to work in a 100-ohm system. The input and output are both matched by a cascade of three L networks. Multiple L networks increase the bandwidth and reduce the circuit losses over that of single L matching networks. For both the input and the output, the inductors for the two L networks closest to the FETs are made from the transistor leads and the PC board traces.

For the input circuit, the highest impedance L network includes capacitors in series with the inductors

parts list for fig. 10

designation	description
C1	6-60 pF mica trimmer, Arco 404
C2,C3	22 pF dipped mica
C4,C5,C10,C11, C12,C13,C14,C15	47 pF Unelco J101 type
C6,C8	100 pF Unelco J101 type
C17	4.7 pF disc ceramic, NPO
C18	2-20 pF mica trimmer, Arco 402
C19	1 µF 35 volt tantalum capacitor
C20,C21,C22, C23,C24,C25	470 pF disc ceramic
C26	1 pF disc ceramic, NPO
C27,C28	6.8 pF disc ceramic, NPO
C29	10 pF disc ceramic, NPO
C30,C31	6.6 pF disc ceramic, NPO
C32	470 pF disc ceramic
C33	51 pF dipped mica
C7,C9,C16	not used
CR1,CR2	1N4148
L1,L2	5 turns No. 20 enamelled wire, 5/32 inch inside diameter
L3	5 turns No. 16 enamelled wire, ¼ inch inside diameter
L4	copper strap 1/8 inch × ¼ inch
L5,L6	2 turns No. 20 enamelled wire, 5/32 inch inside diameter
Q1,Q2	Siliconix DV1260T
R1,R2	10 K, ¼ watt resistor
R3,R4	10 K linear pot
R5	100 ohm, ¼ watt
R6	100 K, ¼ watt resistor
RY1	DC DPDT relay

(C1, L1 and C3, L2). This arrangement is used to break up a low frequency resonance (about **35 MHz**) that can cause oscillations to occur in a push-pull mode. The use of series capacitors is undesirable from the standpoint of loss and bandwidth, but the degradations are not severe. The amplifier outputs are paralleled at less than the 100-ohm point, mainly as a matter of mechanical convenience.

Tuning ease and spectral purity were similar to the

single transistor amplifier. The second harmonic is about 75 dB below the fundamental. An imbalance between the two sides was noticed while tuning, probably because of variations in input capacity. It was found that a 30 pF mica capacitor could be moved along the input lines until the output was maximized. This seemed to be adequate compensation for the imbalance, since after the capacitor was installed the two sides seemed to behave very symmetrically.

temperature effects

The amplifier was also tested in a temperature chamber over the range of −40 to +60 degrees C. Quiescent current, power output, gain, power bandwidth, and spectral purity were all examined. The quiescent current was very slightly temperature dependent. As expected, the current is higher at lower temperatures. The total variation was only 0.4 amperes across the entire range. The room temperature current was set at 6.0 amperes. In terms of RF parameters, the variations were so slight as to be difficult to measure. The gain appeared to be a few tenths of a dB higher at the low temperatures. Current drawn at the 100-watt point did not vary significantly with temperature. In all cases the spectrum was clean. In short, temperature effects are not a problem for this amplifier.

1 1/4-meter amplifier

The final amplifier to be discussed is a 1-1/4-meter amplifier using two paralleled devices. A schematic of it is shown in **fig. 10**. Many of its characteristics are quite similar to those of the lower frequency version, but, as might be expected, both the gain and efficiency were lower at this frequency. For example, with a supply voltage of 16 volts, 10 watts input produced 80 watts output, as compared to 100 watts at the lower frequency. An amplifier designed for 25 watts in and 100 watts out ran at 46 percent efficiency at 13.6 volts, as compared to better than 60 percent at the lower frequency. Quiescent current was set at 6 amperes for both the 2-meter and 1-1/4-meter amplifiers.

conclusion

RF power FETs offer a number of significant advantages over bipolar devices; they are easier to handle and allow somewhat simpler circuits. As manufacturers increase the selection and continue to improve specifications, we will see an increasing number of bipolar and vacuum tube designs being replaced with RF power FETs.

reference

1. J.D. Moell, KØOV, "Forget Memory," *ham radio*, January, 1983, pages 62-64.

ham radio

measuring noise figure

How noise figure, noise factor, ENR, and hot/cold terminations are interrelated

Most Radio Amateurs and electrical engineers learn early in their careers that the ability of a receiving system to respond to weak signals can be improved, not by the addition of gain, but by the reduction of noise.

It is the signal-to-noise ratio that ultimately renders a radio signal useless, and every amplifier or mixer (stage) in a receiver adds to the noise, thus decreasing the signal-to-noise ratio. The amount of noise added by an amplifier or mixer is given quantitatively as a noise factor which is the ratio of the input signal to noise ratio to the output signal-to-noise ratio:

$$noise\ factor = \frac{S_i/N_i}{S_o/N_o} \qquad (1)$$

where S_i = input signal power
N_i = input noise power
S_o = output signal power
N_o = output noise power

If the amplifier frequency response is wide, as compared to the bandwidth of the signal, no amplifier or mixer can improve the signal-to-noise ratio of the input signal. That is, if the input signal-to-noise ratio is 20 dB, the output signal-to-noise ratio will be less than 20 dB. A similar expression, the noise figure, is the "decibel" equivalent of the noise factor and is given by:

$$noise\ figure = 10\ log_{10}\ (noise\ factor) \qquad (2)$$

Thus the signal-to-noise ratio of an input signal is reduced by an amount equal to the noise figure. That is, if a signal with a 20 dB signal-to-noise ratio is passed through an amplifier with a 5 dB noise figure, the output signal-to-noise ratio is 15 dB.

Because the ability of a receiving system is primarily limited by the noise performance of the first RF amplifier, measuring the noise figure is an important part of the receiver evaluation.

noise measurement difficulties

It would appear from the previous equations that measuring the noise figure of an amplifier should be a simple matter of measuring the signal-to-noise ratio of both the input and output of the unit under test. The difficulty in this method arises from the fact that the measuring equipment has noise contributions of its own, and it would be difficult to determine the signal-to-noise ratio of the input signal.

Another problem arises from the fact that the noise power is proportional to the bandwidth used for the measurement. Modulated signals have various relationships between measuring bandwidth and the measured power depending on the form of modulation. Therefore the signal-to-noise ratio can become a rather unpredictable function of the measuring system bandwidths.

This is the electronic equivalent of comparing apples and oranges. The solution lies in making the signal-to-noise measurement using a noise signal, and comparing it to the system noise. Using this method, the measured power is proportional to the measuring bandwidth for *both* signals and noise.

The device used to generate the noise signals is called a noise generator. It is important that the noise power output from a noise generator be greater than the noise power of a resistor at normal room temperature, otherwise the signal-to-noise ratio of the noise generator will be zero dB and of no value for measuring noise figure. The noise power output of any resistor at normal "room" temperature, 290° Kelvin, is:

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$$P_n = KT_oB \quad (3)$$

where P_n = the noise power
B = the measuring bandwidth
K = Boltzman's constant
T_o = 290° Kelvin (17° centigrade)

If the noise power of the noise source is higher than a resistor at 290° K, it is equivalent to the noise power from a resistor at some temperature — say, T', which provides a noise power output of:

$$P_s = KT'B \quad (4)$$

where P_s = the noise power of the noise source
T' = the equivalent noise source temperature

The noise power can be generated in several ways. First, a resistor at the higher temperature or some other artificial method such as using the noise current of a solid-state or thermionic diode could be used to generate the equivalent noise power.

A useful expression for the amount of noise power available from a noise source is called the excess noise ratio, *ENR*, and is given by:

$$ENR = \frac{T' - T_o}{T_o} \quad (5)$$

To measure the noise figure of an amplifier or mixer using a noise source, the amplifier is connected to a power measuring instrument and the noise source is connected. The power output from the device under test is measured. A termination, typically 50 ohms, at a temperature of 290° K is connected to the device under test and again the power output is measured. The ratio of the two power outputs, called the Y-factor, is used to calculate the noise figure. The Y factor is:

$$Y = \frac{P_o'}{P_o} \quad (6)$$

where P_o' is the power output of the device under test with the noise source connected, and P_o is the power output with the 290° K termination. The noise figure of the device under test is:

$$NF = 10 \log_{10} \frac{ENR}{(Y\text{-}1)} \quad (7)$$

In theory, this technique is simple; however, the difficulty lies in making an accurate power measurement of the output power. A resistive termination at 290° K has a power output of −144 dBm in a 1 kHz bandwidth and even after amplification by the device under test, the output is relatively small.

In order to increase the power to a point at which a convenient power meter such as bolometer type may be used, additional amplification is required. The noise figure of the amplifier used after the unit under test (post amplifier) will affect the noise figure measurement, according to the following relationship:

$$NF = NF_1 + \frac{NF_2 - 1}{G_1} + \frac{NF_3 - 1}{G_1G_2} \quad (8)$$

where NF = noise factor of the cascaded system
NF_1 = noise factor of the first amplifier
NF_2 = noise factor of the second amplifier
NF_3 = noise factor of the third amplifier
G_1, G_2 = gains of the first and second amplifiers, respectively

If the noise figure of the post-amplifiers is low or the gain of the unit under test is relatively high, the effect of the post-amplifier noise figure will be small.

If a signal were a simple sine wave, power could be determined by measuring the voltage with a sensitive voltmeter and determining the power by using the simple relationship $P = E^2/R$, where E is the measured voltage and R is the system resistance usually 50 ohms. *Noise* signals are not simple and require a true power measurement. A bolometer type of power meter provides an accurate true-RMS power measurement for noise figure measurements. A diode-type voltmeter will provide the proper response if the diode is operated in the so-called square-law region. This is the region of the point contact diode characteristics where the rectified output voltage is proportional to the input voltage squared. A crystal detector or diode type voltmeter can be checked for square law response by increasing the RF input by 3 dB and checking for a two-to-one increase in the rectified output or, in the case of a voltmeter, a doubling of the meter deflection.

test set-up

Because the square-law region of a diode does not usually cover a large range of power, it is desirable to include an attenuator between the diode detector and the amplifier under test as shown in **fig. 1**.

To make a noise figure measurement using this set-up, the attenuator is adjusted to give a meter reading in the square-law region of the diode with the amplifier terminated (at its input) with a room temperature termination. This can be checked by increasing the attenuator by 3 dB and observing the meter reading.

The room temperature termination is replaced with the noise source (also called a "hot termination"), and the attenuator is adjusted to provide the same output reading as before. The amount of attenuation added is called the Y factor, in dB, which has to be converted to a pure ratio. This is used with **eq. 7** to calculate the noise figure.

The measurement of noise figure requires a known ENR. Most Amateurs optimize the noise figure of a receiving system using a noise diode by adjusting the

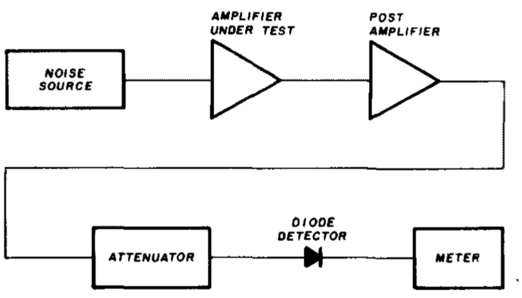

fig. 1. Test set-up for measuring noise figure.

amplifier to give the greatest difference between noise power output with the noise diode on and off. The actual noise figure is not known unless the noise power of the diode is known. Noisy semiconductor diodes do not provide a calculable amount of noise power; commercially available solid-state noise sources are calibrated with a hot-cold noise source.

One very effective method of creating a known ENR is to maintain two identical resistors at a known temperature. Two commonly used temperatures are the temperature of liquid nitrogen at 77° K and the boiling point of water at 100° C or 373° K. The ENR of this system would be:

$$ENR = \frac{T - T_o}{T_o} = \frac{373 - 77}{77} = 3.84$$

or as expressed in decibels: 5.8 dB.

This ENR would be fine for measuring systems with noise figures of a few dB, but it is difficult to maintain the cold termination at 77°K. The cold termination could be at "room temperature" (290°K), and the "hot" termination well above 290°K. One excellent source would be the hot filament of an incandescent lamp. In order to use an incandescent lamp for noise source, the lamp must have a resistance near 50 ohms (hot) and the temperature must be known. Both of these parameters may be determined from the change of resistance versus temperature of tungsten as shown in **fig. 2**.

As an example, assume that a lamp with a cold resistance of 5 ohms is used for a noise source. In order to increase the resistance to 50 ohms, a tenfold increase is required which corresponds to a temperature of 2000° K which produces an ENR of 7.5 dB. This temperature is typical of an incandescent lamp and can be achieved without resorting to special lamps.

A noise source was constructed using two microminiature lamps used for illuminating electronic wristwatches (see **fig. 3**). The lamp is rated for operation at 1.5 volts at 15 mA, which is convenient because this calculates to 100 ohms hot. To determine the temperature of the filament, the cold, 290° K, resistance of the filament was measured and the temperature determined from **fig. 2**. The cold resistance was measured and found to be 16.8 ohms. The hot/cold ratio is 5.9, which corresponds to a temperature of 1300° K.

Two lamps in parallel correspond to a 50 ohm hot source at 1300° K, which is an ENR of:

$$ENR = 10 \log \frac{T - T_o}{T_o}$$

$$= 10 \log \frac{1300 - 290}{290} = 5.42\ dB \qquad (10)$$

Two 1/8 watt 100-ohm resistors were included in the noise source case as the "cold" terminator.

noise source frequency dependence

In order for a noise source to be effective, the impedance must be close to 50 ohms resistive throughout the frequency range of interest. The 100 ohms of the example noise source was determined from the DC operating conditions and does not include the inductive and capacitive components. The complete noise source will have the effects of the coupling

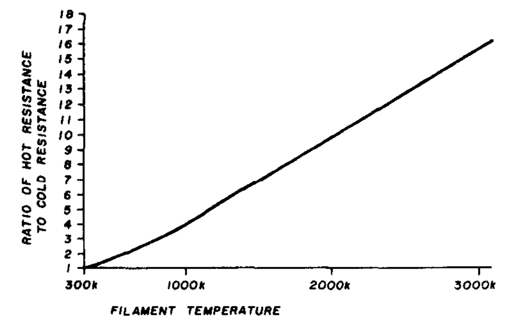

fig. 2. A plot showing the change of resistance with temperature for a tungsten filament.

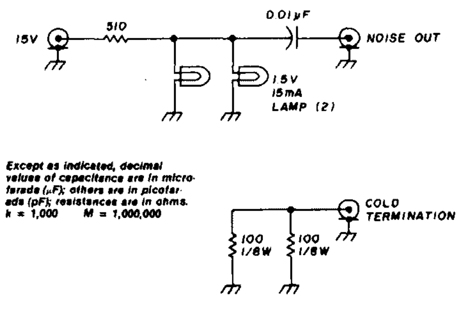

fig. 3. Hot and cold noise sources can consist of 1.5 V incandescent lamps and equivalent 50-ohm resistor, respectively.

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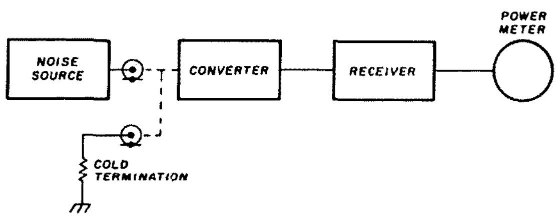

fig. 4. Test set-up for measuring VHF/UHF converter noise figure.

capacitor, the BNC connector and the inductance of the lamp filament. In spite of all of the reactances the return loss as measured on a network analyzer was better than 14 dB up to 1 GHz. Adding a 0.68 pF capacitor improved the return loss to 20 dB to 1 GHz, a very respectable figure for a noise source.

This noise source is suitable for making accurate noise figure measurements to 1 GHz.

The 20 dB return loss of the noise source will allow noise figure measurements to an accuracy better than 0.1 dB.

measuring converter noise

To measure the noise figure of a VHF/UHF converter, the converter is connected to a receiver as shown in **fig. 4**.

An RF probe connected to one of the later IF amplifiers of the receiver is used as a power indicator. Because the noise figure of the system is a function of the setting of the attenuator, the resulting measurement could be distorted. For measuring a noise figure of 2 dB or less with a gain of 20 dB or more, the error will be less than 0.1 dB.

Set the attenuator to 0 dB and terminate the converter with the cold termination. Connecting the converter to the receiver should produce a noticeable increase in background noise. Find a convenient point in the IF amplifier to connect the RF probe. A signal of 100 mW or less is usually in the square law region of the voltmeter. Disconnect the cold termination and connect the hot termination. Adjust the attenuator to provide the same RF voltmeter indications as before; the setting of the attenuator is the Y factor, in dB, and is used in **eq. 7** to calculate the noise factor.

Noise figure measurement determines one of the important operating parameters of receiving systems. The introduction of low-cost GaAs FETs has brought about a reduction of noise figure in many VHF and UHF receiving systems. Using a noise source and the techniques described in this article, it is possible to quantitatively evaluate the improvements effected by the latest technology.

ham radio

verticals over *REAL* ground

Ground system geometry and soil conditions determine performance

How does the geometry of a radial system or ground screen affect the radiation pattern of a vertical? How do ground and roof mounting differ? How does gain change with the frequency of operation? And do reflections depend on the dielectric constant and conductivity of the ground? This article — complementing our series on vertical phased arrays by K2BT — addresses these and other questions, in one of the clearest presentations on this important subject ever to appear in the Amateur literature. — **Editor**

Many hams using verticals have wondered how the ground or earth beyond the radial system affects the radiation patterns of their antennas. Much has been written about the need for a good ground screen or radial system to provide a low-loss return path for ground currents. However, not much has been written for Amateurs about how ground reflections affect the performance of a vertical over real ground.

horizontal versus vertical polarization

For dipoles and other antennas putting out predominantly horizontally polarized waves, reflection from a perfectly conducting ground gives a 180 degree phase shift and reflected electric field intensity equal to the incident field. For real grounds the phase shift remains close to 180 degrees and there is little attenuation of the incident wave. That is why reflection from perfect ground is said to indicate what may be expected of a horizontal antenna over real ground.

The situation is quite different with vertically polarized radiation, which is reflected from perfect ground without phase shift or attenuation. However, only salt water is accurately represented by perfect ground. Any other ground, even saturated, fertilized farmland, provides large phase shifts and attenuations of incident radiation which also strongly depend on the angle of incidence and reflection. At grazing angles of incidence and reflection, the phase shift is 180 degrees and there is no attenuation, just as with horizontally polarized radiation. As the reflection angle increases, the phase shift and reflected field intensity both decrease very rapidly, until an angle is reached where the phase shift is 90 degrees and the reflected field intensity is minimum (maximum attenuation). This angle is called the *pseudo-Brewster angle* and ranges from 1 or 2 degrees for salt water to 30 degrees for rocks or dry sand.[1] At this angle, the radiation pattern of a vertical is affected least by ground reflections. Below the pseudo-Brewster angle, the reflected wave of a ground-mounted vertical partially or wholly cancels the direct wave, while above this angle, ground reflections enhance direct radiation. Generally, the better the ground, the smaller the pseudo-Brewster angle, and the greater the gain from ground reflections.

The radiation pattern of a vertical over infinite, perfectly conducting ground is maximum toward the horizon (takeoff angle of 0 degrees). However, a vertical over real ground has no sky wave radiation

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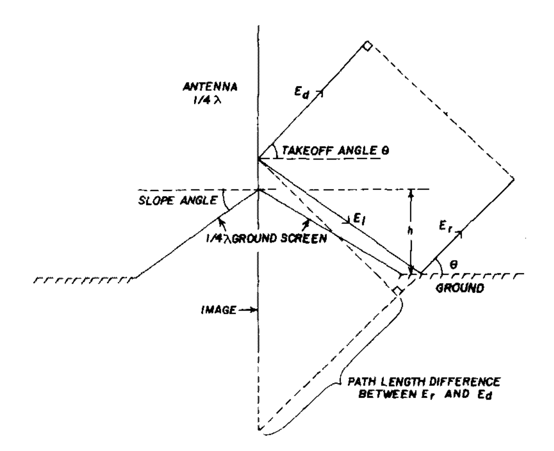

fig. 1. Geometric relationships in a quarter wavelength vertical at height *h* above ground.

toward the horizon and a lobe of maximum radiation varying from 15 degrees, for exceptionally good soil, to 30 degrees, for completely dry rocks and sand. In practice, then, real ground decreases the gain and significantly raises the radiation or takeoff angle of a vertical.

free space patterns and ground reflections

To keep things *simple, we will* concentrate on a quarter wave, current-fed vertical with a quarter wave, perfectly conducting ground plane, mounted over real ground. The radiation patterns of practical verticals with sixty or more radials will closely approach the patterns presented here.

The overall approach is to combine the results of the optical theory of reflection with the free space pattern of a particular vertical configuration. This composite picture shows a radiation pattern modified by reflections from ground having a particular conductivity and dielectric constant.

The free space pattern is calculated from the standard electric field intensity formula for a quarter wave vertical with an infinite, perfectly conducting ground plane.[2] The ground plane defines a collinear quarter wave image "antenna" below the ground screen (**fig. 1**). In other words, a quarter wave vertical with an infinite ground screen is electrically similar to a half wave vertical dipole except in the feedpoint resistance of the vertical, which at 36 ohms, is just half that of the dipole; its power gain is also 3 dB higher. In our case, the ground screen is a quarter wave in radius instead of infinite. Here the image antenna will be electrically shorter than a quarter wave at takeoff angles θ less than some critical angle defined by the geometry of the system. The progressive shortening of the image as the radiation angle is lowered modifies the free space pattern from what you get with an infinite ground screen, in which the image has the same length as the antenna at all radiation angles.

Ground reflections are determined *from the* Fresnel equation for the reflection of vertically polarized electromagnetic waves from a plane surface.[3] A set of reflection coefficients, or ratios of reflected to incident field intensities, are calculated for useful takeoff angles in 5 degree increments over ground of a particular conductivity and dielectric constant. If we call the incident-electric field intensity E_i, the reflected field intensity E_r, and the reflection coefficient R, then $E_r = R \cdot E_i$, and E_r is added vectorially to the free space direct field intensity E_d at each takeoff angle to construct a vertical radiation pattern for a vertical over real ground (**fig. 1**).

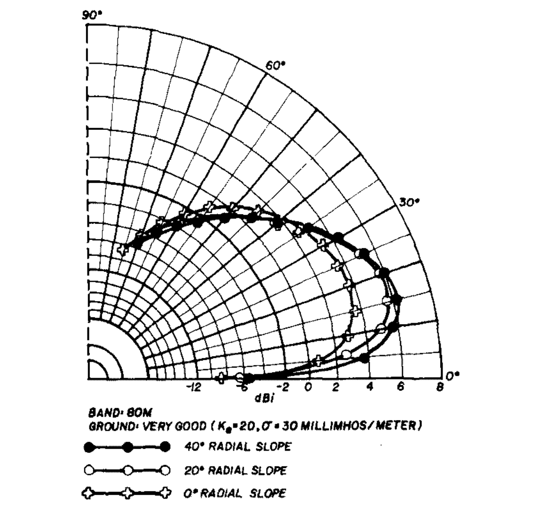

fig. 2. Vertical radiation patterns as a function of the radial slope angle.

ground screen geometry

The radiation pattern of a vertical is strongly influenced by the slope or angle below horizontal of the quarter wave ground screen or radial system. **Fig. 2** shows 80-meter vertical radiation patterns of quarter wave verticals, ground mounted over very good ground (e.g., wet, fertilized farmland). The angles of radial slope are 0 degrees (horizontal radials), 20, and 40 degrees. The configuration with ground systems sloping below horizontal correspond to the antenna mounted atop a mound "carpeted" with a ground screen or dense network of radials. Note that the *verticals with sloping radials have a significant gain advantage over the horizontal or 0 degree radial configuration* — 2.5 dB at a 10 degree

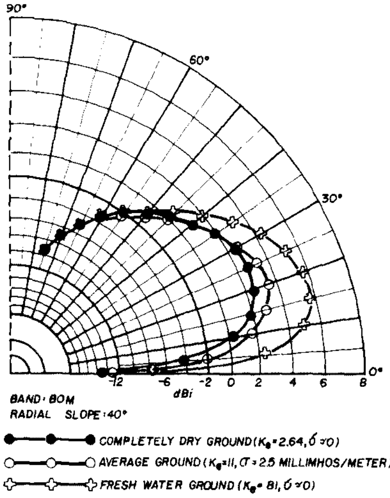

fig. 3. Vertical radiation patterns as a function of ground conditions.

takeoff angle in the case of a 40-degree radial slope. This angle is about the optimum slope. As the slope gets steeper, the gain decreases due to increasing radiation resistance. Other bands besides 80 meters show similar gain increases with sloping radials.

Fig. 3 shows what to expect if you live in a desert or on dry, sandy ground anywhere, or in a city, surrounded by concrete and asphalt. In these settings, the advantage of sloping radials completely disappears. Even over average ground with moderate dielectric constant and conductivity, little is to be gained from a sloping ground screen. Over average or poor ground, the combined phase shift from reflection and from the path length difference between E_i and E_r (**fig. 1**), is between 90 and 180 degrees for useful takeoff angles; this large phase shift leads to substantial cancellation of E_d by E_r.

The advantage of a sloping ground screen is regained over a fresh water ground, even though the conductivity is insignificant (**fig. 3**). Compared to very poor ground, a fresh water ground offers a gain of up to 7 dB and compared to average ground, a gain of 4 dB. This is because the high dielectric constant of water provides a good reflecting plane.

These radiation patterns suggest that it's possible to enjoy a seasonal advantage with sloping radials. Several days of rainy weather can raise both the dielectric constant and the conductivity a great deal, leading to another 2 to 6 dB of reflection gain at takeoff angles below 30 degrees.

roof or tower mounting

Unless a vertical is mounted two to three wavelengths high, putting it up in the air does not increase the gain, as it does with most horizontal antennas. In fact, about 1 dB at one quarter wavelength high is lost, compared with a ground-mounted vertical (**fig. 4**). Of course, if your ground-mounted antenna is completely boxed in by cars, buildings, or other obstructions, getting the current loop above the lossy surroundings may still pay off. The gain trends with radial slope seen for ground mounted verticals also appear for elevated verticals.

About half a wavelength above ground at the base is a particularly unfortunate height at which to mount a vertical over most grounds. **Fig. 4** shows the 40-meter radiation pattern of a quarter wave vertical

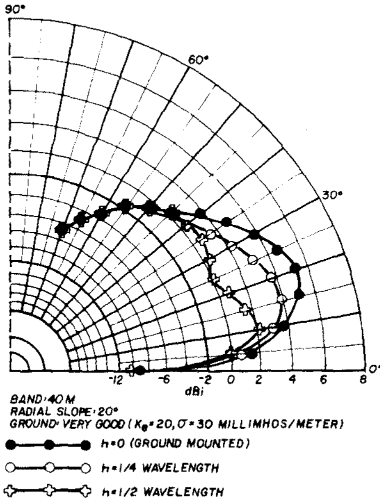

fig. 4. Vertical radiation patterns of a vertical mounted at various heights above ground.

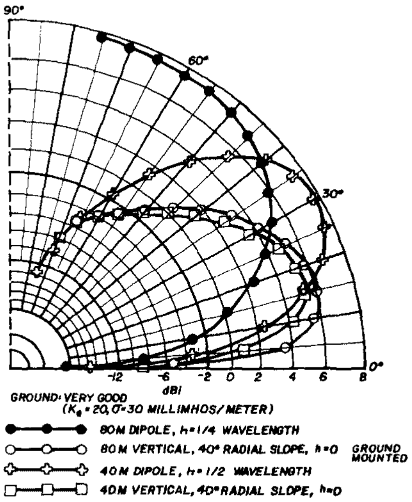

fig. 5. Vertical radiation patterns of 40 and 80-meter verticals and horizontal dipoles.

with 20° radial slope, at one half wavelength high. It has up to 5 dB less gain at the most useful radiation angles than the same antenna, ground-mounted over very good ground. Other radial geometries at one-half wavelength show similar losses. Only a vertical over salt water shines at this height, giving about 7 dBi gain at 5 degrees, which is especially good for the bands above 30 meters.

vertical versus dipole

Recently there has been interest in discussing whether the vertical or dipole works better on the various bands.[4] **Fig. 5** compares the 80-meter radiation pattern of a ground-mounted quarter wave vertical having 40-degree ground screen slope, with the broadside pattern of a half wave dipole one quarter wavelength above very good ground. The dipole at about 65 feet (20 meters) is as high as most Amateurs would mount an 80-meter dipole. The vertical shines at takeoff angles below 30 degrees, while the dipole takes over at higher angles. Undoubtedly some 80-meter DX comes in at angles above 30 degrees, as do most short-skip signals. In that case the broadside dipole looks like the better antenna. However, for most really long-hop circuits, the vertical would be the better choice.

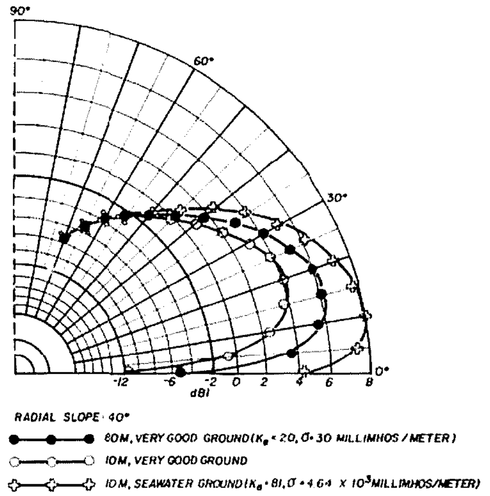

fig. 6. Vertical radiation patterns of 10 and 80-meter verticals over differing grounds.

calculating vertical radiation patterns

To calculate vertical radiation patterns, first define the following terms:

E_d direct (free space) field intensity
E_i field intensity incident on the ground
E_r field intensity reflected from the ground
R reflection coefficient, equal to E_r/E_i
K_e dielectric constant of the ground
σ conductivity of the ground in millimhos/meter
λ wavelength in meters
θ takeoff angle of the direct ray. Also equal to the angle of incidence and reflection (**fig. 1**)
ϕ angle of refraction of the transmitted ray
ρ phase shift due to reflection
β phase shift due to the path length difference between direct and reflected rays (**fig. 1**)
h height of the antenna above ground, in meters
G_i gain in dBi relative to an isotropic radiator

The electric field intensity E_d or E_r is expressed as a function of the current, which is a standing wave on the antenna (radiator plus image). This current function is integrated over the length of the antenna, including the image, to give the free-space field intensity at a particular θ. If $\theta > 0$, the intensity is E_d and if $\theta < 0$, we call the intensity E_i. E_r corresponds to the ray reflected from the ground.

R is calculated from the Fresnel equation for reflection of vertically polarized radiation from a plane surface:

$$R = \frac{\sqrt{K_e'}\,sin\theta - cos\phi}{\sqrt{K_e'}\,sin\theta + cos\phi}$$

where $\phi = sin^{-1} \frac{1}{Re\sqrt{K_e'}} cos\,\theta$ from Snell's law, and the complex dielectric constant

$$K_e' = K_e - j(6 \times 10^{-2}) \cdot \sigma \cdot \lambda$$

For each θ, E_d is calculated and for $-\theta$, E_i, the corresponding ray directed downward, is calculated. Then $E_r = R \cdot E_i$ and E_t is determined by adding the E_d and E_r vectors:

$$E_t = \left[E_d^2 + 2E_dE_r\,cos\,(\rho + \beta) + E_r^2\right]^{1/2}$$

The gain at a particular takeoff angle is $G_i = 20\,log\,E_t + A$, where A is a normalization factor which gives the correct gain of the vertical being considered at 0 degrees takeoff angle over infinite, perfectly conducting ground. For example, $A = 5.16\ dBi$ for a quarter wave vertical with 0 degrees radial slope at ground level over infinite, perfect ground. Thus, various combinations of radial slopes and grounds will all give G_i expressed in dBi.

On 40 meters, a half wave dipole one half wavelength above very good ground, pitted against a ground-mounted vertical with 40° radial slope, has a broadside gain advantage above 15 degrees (**fig. 5**). This angle is about the median for New England to western European circuits on 40 meters.[5] Again, the dipole looks better for most medium and short-skip circuits, while the vertical has a slight advantage for very long hops and band conditions favoring very low radiation angles.

On 10 through 30 meters, a vertical is at more of a disadvantage because its reflection gain drops off faster with increasing frequency than the reflection gain of a dipole. As **fig. 6** shows, on 10 meters a ground-mounted vertical with 40 degree radial slope over very good ground has 3 dB less gain at a takeoff angle of 10 degrees and 4 dB less gain at 5 degrees than a similar vertical on 80 meters. A vertical on 10 through 30 meters simply can't compete with a dipole one or more wavelengths high over any ground except salt water. *Between 30 and 40 meters is the point at which the dipole really takes over.*

conclusion

No doubt the long-standing debate will continue over whether a vertical or a dipole is the better antenna. I hope the results presented here will encourage more good A-B comparisons of verticals with other simple antennas. The worldwide system of beacons on 14.100 MHz should aid in these comparisons.

These results also show that ground mounting in the clear is generally the best way to set up a vertical. If the ground beyond the radial system is exceptionally good, the radials can be sloped downward 20 to 40 degrees for another 2 dB or more of reflection gain. Over salt water, a vertical with ground system sloping down to water level and 0.5 miles (0.8 km) of salt water in all directions would be a hard antenna to beat, especially on 10 through 30 meters, where 7 dBi of gain below 5 degrees is what is needed for long hop DX (**fig. 6**). However, for most grounds the ability of the vertical to launch low angle (below 15 degrees) radiation has often been overrated in the Amateur literature.

acknowledgement

I thank Dick Coombe, K9VPK, for his interest and encouragement throughout this project.

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4. Bill Orr, "Ham Radio Techniques," *ham radio*, October, 1982, page 20; February, 1983, page 79.
5. *ARRL Antenna Book*, 14th Edition, American Radio Relay League, Newington, Connecticut, page 1-10.

ham radio

VHF/UHF WORLD

Joe Reisert W1JR

the VHF/UHF challenge

Although I'm usually branded as a VHF/UHF person, I'm really a DX'er and go where the DX goes, be it 160, 20 meters, or 13 cm. (I'm fully equipped for 13 different Amateur bands and operate them all!) So after I made the sought-after DXCC *Honor* Roll in 1968, I decided that someday I'd probably have every country confirmed on HF (which I did in 1980) and that the best challenges remaining would be on the VHF/UHF bands. I never abandoned the HF bands — as any DX'er will tell you — but I now tend to concentrate my efforts on improving the state-of-the-art on the frequencies above 50 MHz.

a review

It wasn't too many years ago that any Amateur Radio operator venturing above 29.7 MHz had to be ready to "roll his own" rig, so to speak. That's different now. For those so inclined, there has been vast improvement in components and antenna design. You can now purchase commercial equipment for all Amateur frequencies up through 1300 MHz, even to the limits of legal power. This has even been extended to OSCAR and EME (Earth-Moon-Earth). But herein lies the problem: what do you build or buy, how do you hook it up, how do you use it, and more importantly, how do you get the most out of your equipment and keep up with the state-of-the-art?

opportunities

Let's digress briefly and discuss some of the opportunities awaiting us on the VHF/UHF frequencies. This is truly the area for propagation research, and there are records galore just waiting to be made or broken. Disregarding EME and OSCAR for the moment, look at what's happened on 6 meters in the last few years: WAC has been worked by many. Even places like India, Gambia, the Galapagos Islands, Cyprus, and other equally distant places have become workable from North America via F2 propagation.

Then 2 meters opened on the transequatorial path.[1] To our surprise, it wasn't as spotty as expected. In fact, 220 MHz QSOs were finally made and a one-way 70 cm QSO has been verified. Speculation on the existence of a new propagation mode was heard, with hints that FAI (Field Aligned Irregularities) similar to the transequatorial path could possibly exist in the mid-latitudes such as the U.S.A. This was finally confirmed, and QSOs on 2 meters using this mode have been accomplished numerous times; but while it's still possible above 148 MHz, this has yet to be done!

A true Amateur communications satellite, OSCAR 10, is now available to all Amateurs. Then there's EME, the ultimate challenge on Amateur power levels. Worldwide QSOs now take place daily on frequencies from 144 to 2320 MHz. These QSOs are no longer the result of a concerted effort by many individuals pooling their efforts; instead, they're being done by the everyday Amateur who may live on a 60 by 100-foot lot, but is willing to take on the challenge, do the research, and build up a good station. And while this isn't a task for the timid, many non-technical persons are doing it every day.

Let's look beyond operating and propagation. These discoveries weren't made by accident; *they happened and were exploited because the state-of-the-art gear had been improved.*

benefits

So you say, what's in it for me? *PLENTY*. You too can contribute to the state-of-the-art. There is much to be done: some may contribute labor, helping assemble large antenna arrays; others may assist with parts procurement, machining, research, scheduling, antenna design, receiver design, transmitter design and so on. But there's something here for everyone, and the rewards are numerous: records to be broken, propagation modes to be discovered (this takes time and scheduling), new improved antenna systems *(better mechanical designs as well as electrical parameters)*, better receivers with lower noise figure and high dynamic range *at the same time,* and more efficient and cleaner transmitters. (For you HF'ers, have you ever tried to use a frequency adjacent to a station running 100,000 to 1,000,000 watts effective radiated power?) The satisfaction will be enormous. The VHF/UHF frequencies offer a great test bed where you can develop new techniques, circuits, and antennas even if you are restricted to a small or not-so-great location.

summary

Why all the hype? I'm constantly confronted by very competent individuals who are willing to build or buy their own gear to get on the VHF/UHF frequencies but always ask the same questions: "What's the best preamplifier circuit? What's the best antenna for my application? What band should I try?" The answers are not always simple — and often depend on the individual — but there is always one problem common to all the inquiries: where do I find the information I need to put together this gear? The answer is *right here*, because I've been asked by Rich Rosen, the Editor-in-Chief of *ham radio*, to launch this column as a continuing series. Over the coming months I'll try to discuss many of the questions most commonly asked by the newcomer as well as the seasoned VHF/UHF'er. This will be done with concise descriptions, charts, and circuits. After establishing a solid base of factual material over the first few months, I'll then attempt to build on this basis by offering updates, new or state-of-the-art equipment design information, advice on construction of antennas, propagation information and discussion of other subjects that will keep us all up to date. Time and space permitting, we may even include material or ideas expressed by others. Each month I'll also try to list events that are of special interest to VHF/UHF'ers.

Does this seem like the kind of information you're looking for? If so, I'll see you next month.

references

1. Joseph H. Reisert, W1JR, and Gene Pfeffer, KØJHH, "A Newly Discovered Mode of VHF Propagation," *QST*, October, 1978, pages 11-14.

VHF/UHF coming events

Quadrantides Meteor Shower: Predicted peak at 0100 UTC on January 4, 1984.
Best EME weekend: January 20, 21, 1984.
ARRL VHF Sweepstakes Contest: January 14 - 16, 1984. (See December, 1983 *QST* for further information.)

ham radio

meet Joe Reisert . . .

First licensed in 1951 as WN2HQL, Joe Reisert earned his Extra Class license in 1956. He has since held the following call signs: WA6TGY, W6FZJ, W1JAA, and now, W1JR. He attained the DXCC Honor Roll in 1968 and presently has 357/315 confirmed. Joe's interest in UHF began in the late 1960's and he made his first 432 MHz contact in 1970; he has since worked all states on 432 MHz and *nine* other bands: 160 through 2 meters (160, 80, 40, 30, 20, 15, 10, 6, and 2 meters). Active in EME operation on 144 through 1296 MHz, he is the former joint holder of the 2304-MHz tropo record of 330 miles set in February, 1974.

Joe's diversified technical interests are best illustrated by his publication of over thirty articles on subjects as varied as the following: a wide-band, low-noise preamplifier; VHF antenna arrays for high performance; a 432-MHz kW stripline amplifier modification; a low-noise VHF/UHF receiver design; a 2-meter transmitter filter for mode J; new modes of VHF propagation; and the PROMIS Microwave data and video link.

***ham radio* readers can look forward to reading about these subjects and more as Joe Reisert's column continues over the coming months. Welcome aboard, Joe!**

K2RR

G.O.E.S. reception: a simple approach

Inexpensive surplus components and scope bring you satellite weather pictures

Too often Amateurs will not attempt a project because it appears, at first glance, to be far too complex; reception of weather images from the geosynchronous satellites is certainly one such project. After reading several excellent articles on the subject,[1-3] I was convinced I'd need at least a four to six-foot dish and a $400 downconverter in order to build the required equipment. But I decided to try to build the system anyway, using whatever surplus materials I could obtain.

antenna

While I did not — and still do not — have a four or six-foot microwave dish antenna, I did have a small 30-inch dish that had been collecting dust for about fifteen years. Now, the gain difference between a six-foot dish and a thirty-inch dish is determined by the ratio of the dish areas. The six-foot dish has an area 5.76 times the thirty-inch antenna. This translates to a gain difference of 7.6 dB. The focal length of the thirty-inch dish was measured to be 10.8 inches. (The focal length of an existing dish is easily found by dividing the diameter squared by sixteen times the depth.[4]) Next, the F number, or focal ratio, was calculated. This is simply the focal length divided by the diameter, just as for photographic lenses. The focal ratio for the thirty-inch dish is 0.36.

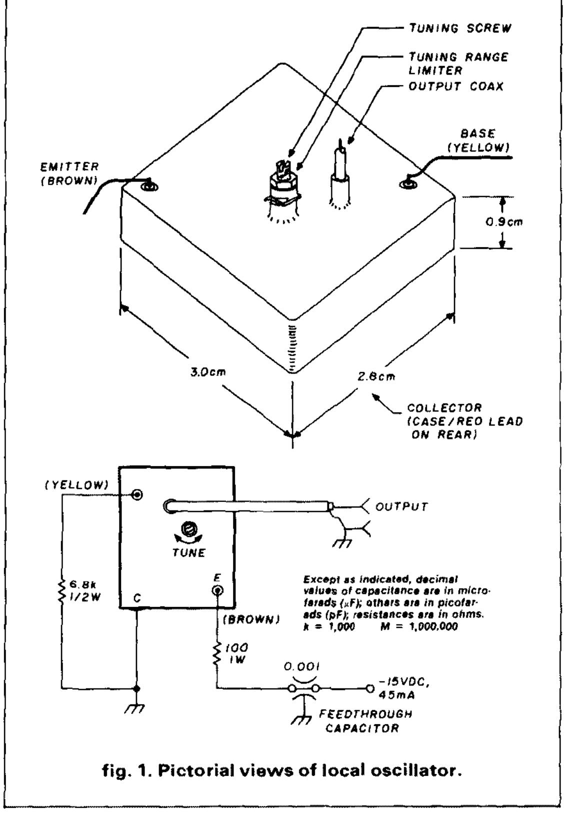


fig. 1. Pictorial views of local oscillator.

By John M. Franke, WA4WDL, 1310 Bolling Avenue, Norfolk, Virginia 23508

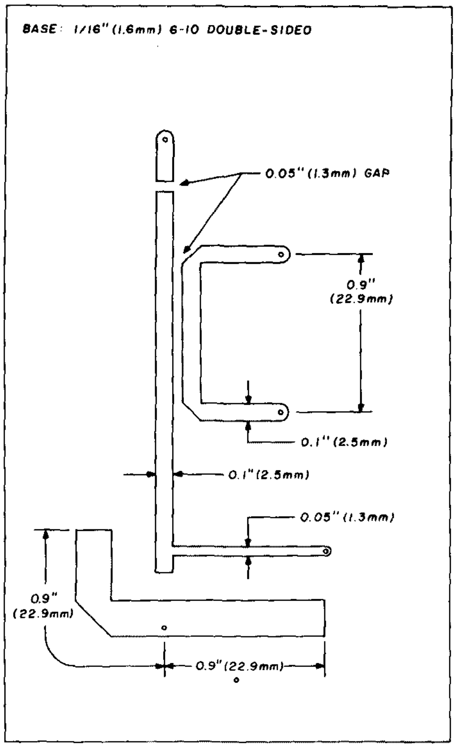


fig. 2A. Mixer printed circuit board.

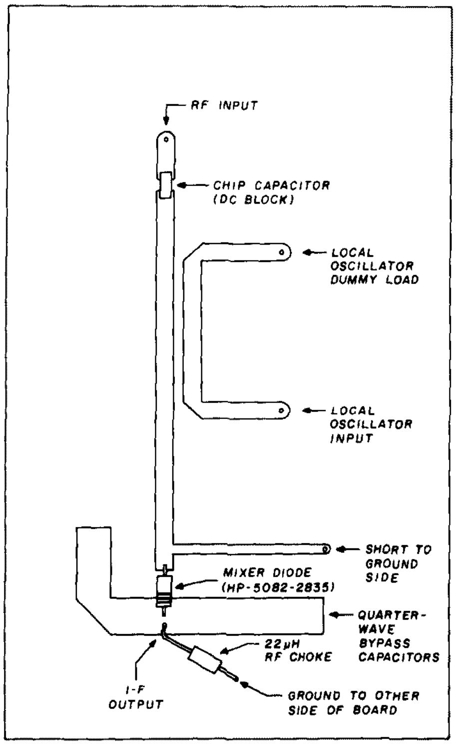


fig. 2B. Mixer components placement.

Once the dish was characterized, I designed the feed, selecting a horn feed because it was the simplest. (After all, this is an Amateur project, not a research program.) I used a piece of five-inch outside diameter aluminum tubing, following an excellent article on waveguide or horn feeds for dish antennas.[5] I chose RG-58 for the feedline from the antenna to the preamp.

antenna mount

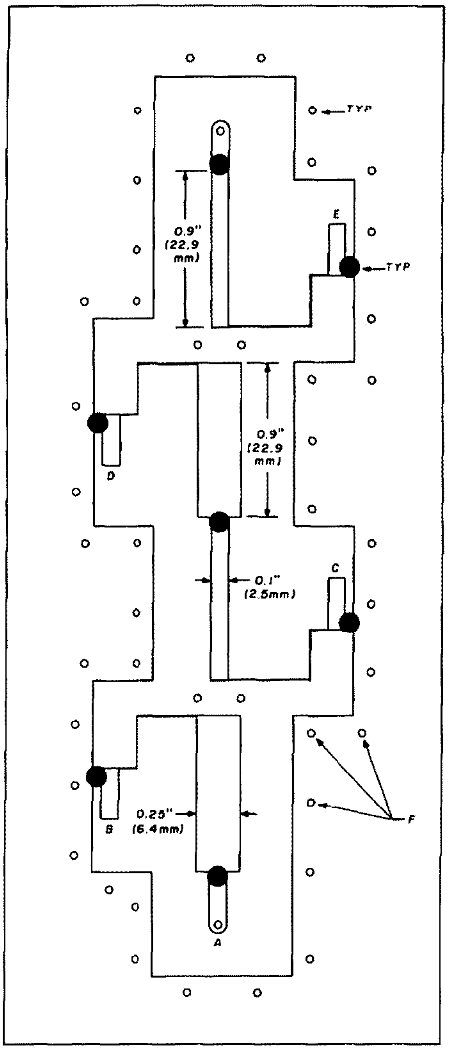
The mount consists of a simple frame made from rack panels and aluminum angle supports. The elevation and azimuth angles for the G.O.E.S.-East satellite were calculated for my station latitude and longitude.[6]

The dish is aligned in azimuth by simply turning the entire mount. A simpler, more compact, mount (see photograph) would be hard to devise. The mount was initially aligned with a compass taking magnetic declination into account. The beamwidth of the antenna is about sixteen degrees; hence, alignment is not critical. The feed can be rotated to adjust for optimum polarization. GOES-East and GOES-West satellites are vertically polarized and GOES-Central is horizontally polarized.

downconverter

The downconverter consists of a mixer and a local oscillator. The local oscillator is a solid-state unit from a weather balloon telemetry transmitter and was purchased at a local hamfest for five dollars. The antenna and modulator were removed and an output connector and bias network added (see **fig. 1**). Though the solid-state unit is difficult to obtain, Fair Radio Sales still carries a vacuum tube version for under five dollars. Output power from either source is high: 20 to 100 mW. The output frequency is nominally 1680 MHz ±20 MHz with a *manual* tuning control.

The mixer uses a single-ended diode with the local oscillator signal inserted using a directional coupler. **Fig. 2** shows the dimensions used with 1/16-inch glass epoxy double-sided printed circuit board material. One side is unetched. The etched side was laid out by first covering the entire copper surface with transparent tape, using a razor blade to trim away the unwanted areas. The areas to be etched away were left covered. (While this is the reverse of most tape techniques, my reasoning will soon become clear.) The board is now sprayed with any color lacquer that is handy; I prefer flat black. The remain-



● DENOTES LOCATION OF CHIP CAPACITOR (7 PLACES)
A RF INPUT
B BASE OF FIRST STAGE
C COLLECTOR OF FIRST STAGE
D BASE OF SECOND STAGE
E COLLECTOR OF SECOND STAGE
F SHORTS TO CONNECT UPPER AND LOWER COPPER PLANES (39 PLACES)

fig. 3. Two-stage preamplifier printed circuit board.

	1st STAGE	2nd STAGE
R1	2.2k	2.2k
R2	4.7k	4.7k
R3	680	560
R4	1k	1k

Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms.
k = 1,000 M = 1,000,000

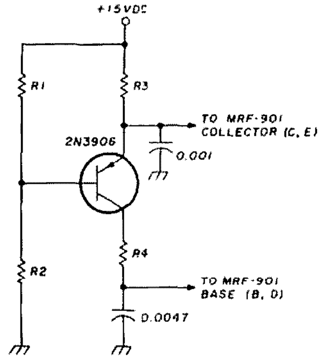


fig. 4. Active bias circuit for MRF-901 preamplifier.

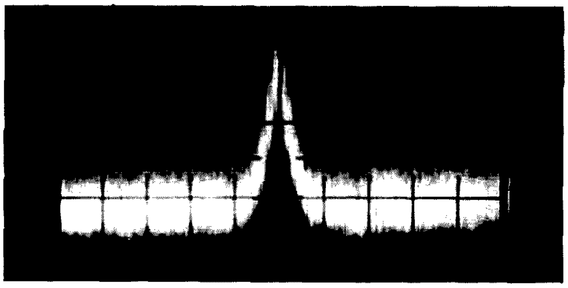


fig. 5. Satellite signal on ART-26 panadapter. Signal-to-noise ratio is approximately 7 dB.

ing tape is peeled off carefully. The unwanted areas are now bare and the desired areas are protected by lacquer. The ground side of the board is painted. After drying, the board is etched with ferric chloride.

The finished board is scrubbed with steel wool to remove the paint and wiring is completed. Care must be taken when installing the connectors to make sure that copper is removed from around the center pinhole on the groundplane side to prevent shorts. (I used SMA connectors, but BNC connectors would also work.) Connectors are placed at each end of the local oscillator coupling line. The end nearest the mixer is the input; a terminator or dummy load is connected to the most distant connector. About 750 microamperes mixer current is normal. This can be measured by lifting the ground end of the choke and inserting a ten-ohm resistor. Measure the voltage drop across the resistor and calculate the current. Too high a resistance reduces the mixer current, giving a false answer. The noise figure is probably around 12 dB.

intermediate amplifier

The IF frequency I used is 30 MHz. This frequency was chosen for two reasons. First, it fell within the range of the difference between the tunable local oscillator and the 1691 MHz signal. Second, it is the center frequency for my ART-26 panadapter, an invaluable tuning aid. (The panadapter cost only five dollars at the same hamfest where I purchased the local oscillator.) In fact, the panadapter was used as my first demodulator by reducing the scan to zero and using slope detection. The IF amplifier used has

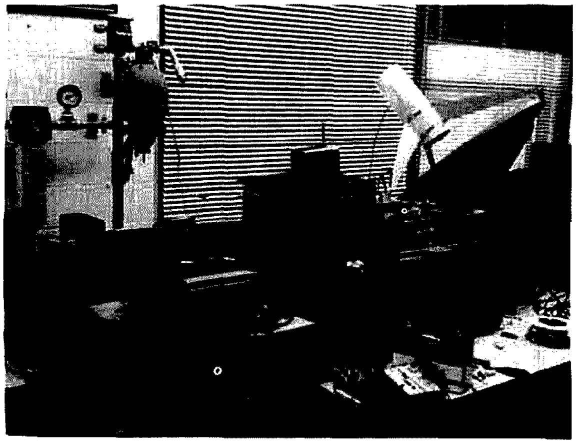


fig. 6. Desktop G.O.E.S. receiver. Window blinds cause no loss.

a gain of 60 dB. Any surplus 30 MHz radar IF could be used. A minimum bandwidth of 30 kHz is needed — more is preferred (2 MHz) if a panadapter is used.

detector

I do not have a 30 MHz FM discriminator, so I up-converted the 30 MHz IF to the FM band and used an FM monitor receiver for detection.

preamplifier

Several excellent articles on low-noise amplifiers are available.[7-9] I decided to build a two-stage amplifier, designing around the MRF-901 transistor. The layout is shown in **fig. 3**. The heavy, wide lines are quarter wave transformers and the narrow lines are quarter wave chokes. The board is laid out using tape and spray paint in a manner similar to that used for preparation of the mixer unit. The dual emitter leads pass through the board and are soldered to the groundplane. Because the emitters are grounded, active baisin̂g is used. The circuit in **fig. 4** was wired on a small printed circuit board and mounted in the preamp housing. Short leads prevent oscillation. The PNP biasing transistors are wired as constant current generators. The collector current of the MRF-901 is determined by the bias transistor emitter resistor and the collector voltage of the MRF-901 is determined by the base voltage of the bias transistor.

100 pF chip capacitors are used for coupling and bypassing. The overall gain is about 18 dB and the noise figure is better than that of the converter operated alone.

The total cost of the project at this point is about $18.00.

In the photograph of the panadapter (**fig. 5**), the signal-to-noise ratio can be seen to be about 7 dB. If the antenna were set up outside, the signal-to-noise ratio would be 10 dB.

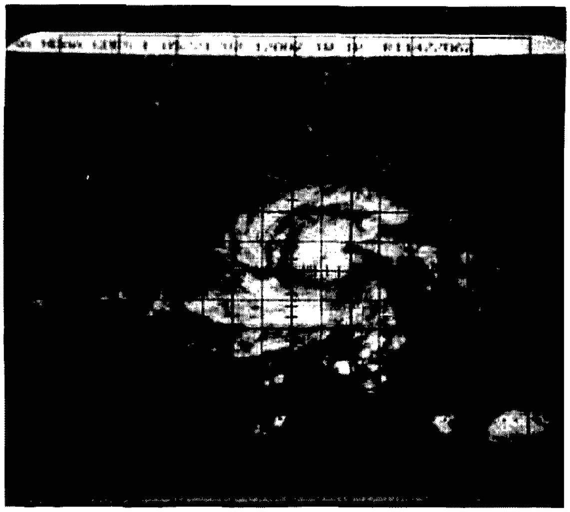


fig. 7. Infrared view of tropical storm "Adolph," May 21, 1983.

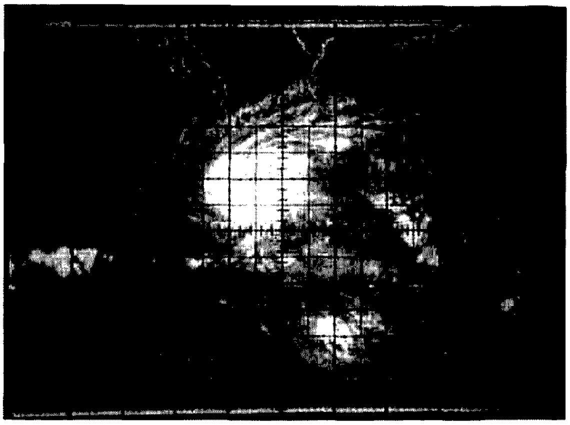


fig. 8. Infrared view of tropical storm "Adolph," May 23, 1983.

display

To photograph the satellite signals, I borrowed a Tektronix oscilloscope and camera (**fig. 6**), using Polaroid type 667 film (ASA 3000). The photographs (see **figs. 7, 8,** and **9**) were recorded in real time. The horizontal sweep was triggered by a 4 Hz signal from a color burst crystal oscillator. A variable divider chain was used for initial phasing. The vertical sweep was generated with a homebrew 10 bit D to A (digital-to-analog) converter (**fig. 10**), connected to a counter that counted horizontal sweep pulses. The video was fed, unfiltered and unrectified, to the oscilloscope Z axis.

conclusion

Future work will include moving the antenna out-

side, possibly building a four-foot dish, tape recording the signal and replacing the mixer/local oscillator with an interdigital crystal-controlled unit. In my experience, I've found that the best way to get involved with microwave technology is to use what you have and get on with it; don't be put off by high prices or sophisticated equipment. (Please enclose an SASE with any questions.)

fig. 9. Visible light view of tropical storm "Adolph," May 23, 1983.

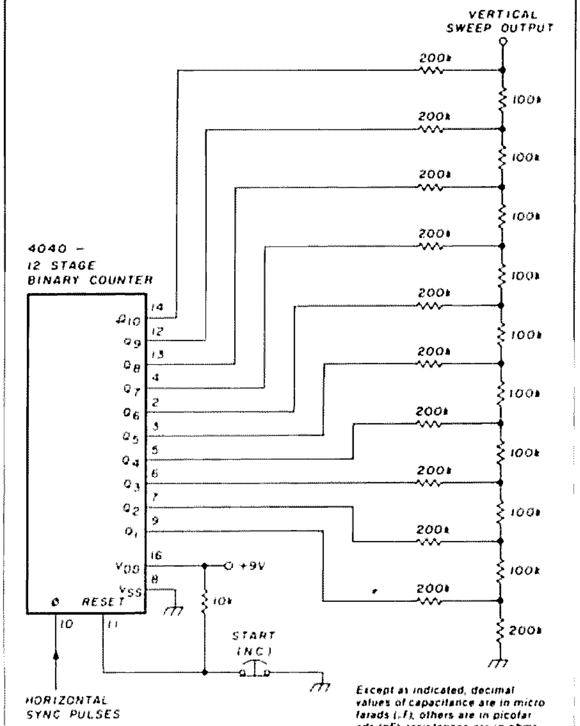


fig. 10. Vertical sweep generator schematic (10-bit Digital to Analog converter).

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ham radio

a wide-range ohmmeter

Measure values from 0.1 ohms to 1000 megohms on *seven* ranges

The typical VOM or digital multimeter found on most workbenches simply cannot handle the extremes of resistance values most Amateurs encounter in building or repairing sophisticated electronic equipment. The usual upper limit of 20 megohms falls far short of required measurements; the usual lower limits could also be extended.

Because I needed an ohmmeter with a wider range than the one on my workbench, I decided to design and build one. But wide range would not be enough; my meter would have to be low in cost, provide relatively good accuracy, be easy to read, and offer reliable long-term calibration. Low cost was achieved by developing a simple circuit and using some junkbox components; accuracy, by including an analog instrument; easy reading, by using a single scale on a 4-1/2 inch meter, thus avoiding incorrect readings sometimes made with multiple scales; and long-term calibration, by use of a line-operated power supply instead of batteries.

I wanted an upper limit of 1000 megohms and a lower limit of 0.1 ohms. Although the least significant bit of a 3-1/2 digit DMM displays increments of 0.1 ohms, the actual resistance could be anywhere between 0.06 and 0.14 ohms for a 0.1 ohm display. On some DMMs the LSB will jump up or down one count which, of course, invalidates very low value measurements.

how it works

The circuit shown in **fig. 1** employs a regulated low voltage, 1.22 volts, which is applied to the unknown resistor in series with a switch-selected standard resistor. The voltage drop across the unknown is fed to the input of one section of dual op amp, LF353. This section is connected as a voltage follower. It has the very high input resistance of 10^{12} ohms, which hardly loads down the unknown. Output null is obtained with adjustment of R18. The next stage, the other section of the LF353, provides an adjustable gain of a little more than 1 so that the meter can be calibrated to read full-scale with the unknown terminals open. When an unknown resistor is connected across the terminals, the meter reads down scale, reaching zero when the unknown resistor terminals, $\pm$ Rx, are shorted together at the panel.

circuit details

The power supply, using a full-wave bridge rectifier and center-tapped transformer secondary, provides zener-regulated +5 and −5 volts for the LF353. An LM317T voltage regulator supplies the

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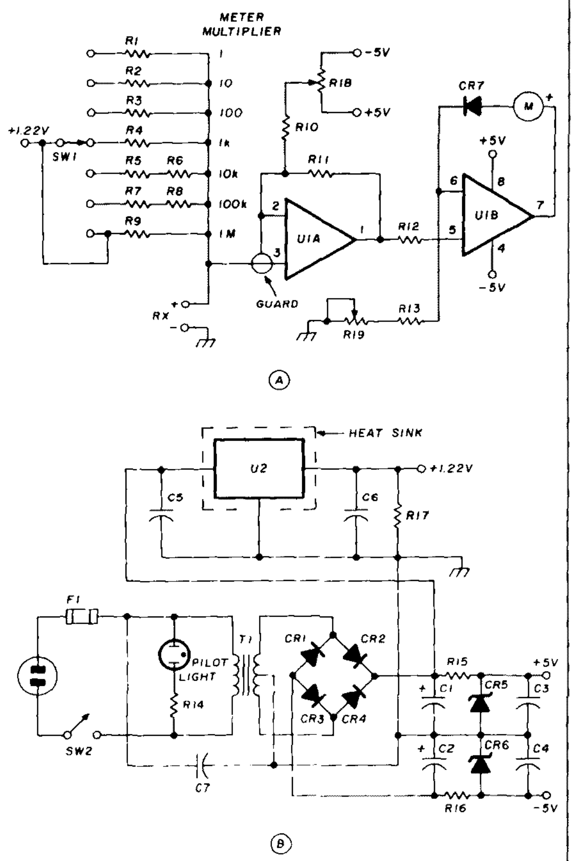


fig. 1. Wide range ohmmeter schematic A) measuring circuit, and B) power supply.

1.22 volts required for the input divider. A heat sink is used on this regulator because of the wattage dissipation on the low range. On this range, with the unknown resistor terminals shorted, 122 mA flow through the 10-ohm standard resistor. Since about 7.5 volts is dropped across the regulator, the worst-case dissipation is 7.5 volts × 122 mA or about 0.9 watts. R17 provides a minimum load of 25 mA for stable operation of the regulator.

The 10-megohm standard is paralleled across the other range standards for all switch positions, which assures an input ground return for the LF353 at all times when switching ranges. Series resistors are added to the × **10k** and × **100k** ranges to compensate for the 10 megohm shunting. The 1N914 diode in series with the meter prevents reverse deflections of the meter. The standard resistors should be metal film type of ±1 percent tolerance.

The null circuitry is conventional for voltage followers. For the values shown, ±4 mV maximum is developed across R11. If more null compensation is necessary with the particular LF353 used, simply increase the value of R11. The internal resistance of the meter is not critical since it is in the feedback loop. Mine had a resistance of 1200 ohms.

On the high end of range × **1M**, it is important to minimize any leakage resistance across the unknown terminals, ±Rx, because such leakage would be in parallel with the resistor being measured. The schematic shows only the range resistors and pin 3 of the first LF353 connected to the positive Rx terminal. Since the common ends of the range resistors are not mounted on any terminal strip supports, but rather supported by their leads, only negligible leakage to ground exists. Consequently, pin 3 exhibits a 10^{12} ohm resistance to ground and thereby adds an insignificant shunting effect. Not shown, however, is PC board surface leakage from pin 3 to ground, if pin 3 were soldered (via a socket) to a PC board trace. PC board surface leakage can be minimized by using a guarding technique. This technique consists of a

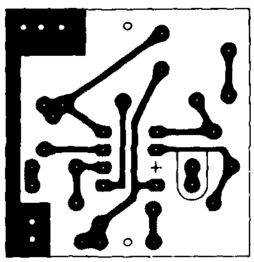


fig. 2. PC board artwork, foil side. Cross indicates drilling location for pin 3 guard terminal. See text.

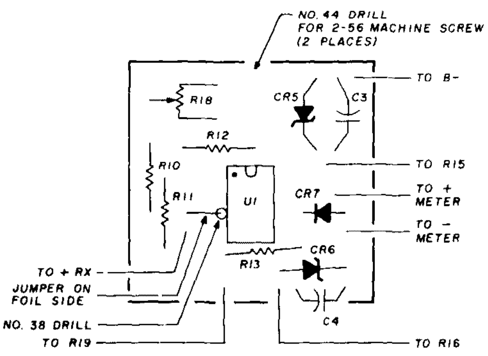
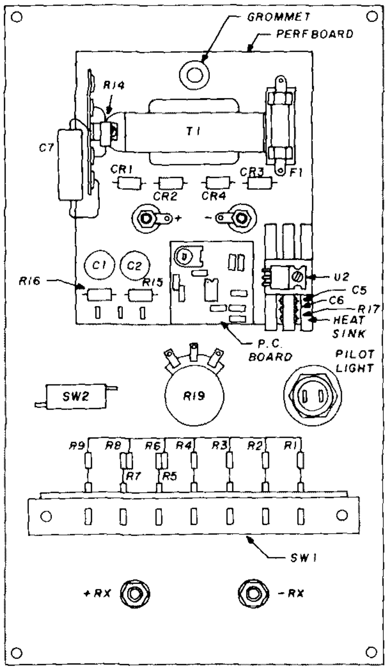


fig. 3. Component side of PC board.

trace ring that completely circles pin 3 and returns it to a low impedance point of voltage — equal to the voltage on pin 3. This must be done without pin 3 (actually pin 3 of the socket) touching the PC board. Isolation is accomplished by drilling a clearance hole in the PC board so that the socket pin 3 protrudes through the board clearance hole and, with a jumper wire, is connected to the trace pad inside the guard ring. The guard ring is connected to pin 2. Purists will be quick to point out that the guard ring should be connected to pin 1 because pin 2 has a higher impedance than pin 1. They are correct; this is only a quasi-guard solution, but for this application, it is sufficient.



parts list

C1,C2	500 μF, 15V aluminum electrolytic P C board terminals
C3,C4,C5,C6	0.1 μF, 50V ceramic, axial leads
C7	0.047 μF, 600V Mylar
CR1,CR2,CR3,CR4	1N4001 50V, 1A rectifiers
CR5,CR6	5.1V zeners, 1N751, 400 mW
CR7	1N914 diode
U1A,B	LF353N duel BiFet op amp, Radio Shack 276-1715
U2	LM317T 1.22v regulator, Radio Shack 276-1778
F1	1A fuse and holder
M	100 μA meter, 4-1/2" square
P.L.	neon pilot light, NE51 with assembly
resistors: 1/4 watt, 1% tolerance metel film	
R1	10 ohms
P2	100 ohms
R3,R6	1 kilohm
R4	10 kilohms
R5	100 kilohms
R7	1 megohm
R8	110 kilohms
R9	10 megohms
resistors: 1/4 watt, 5% tolerance carbon composition	
R10	2.2 megohms
R11	1.8 kilohms
R12	4.7 kilohms
R13	10 kilohms
R14	220 kilohms
R15	390 kilohms
R16	470 kilohms
R17	47 ohms
R18	25 kilohm trimmer potentiometer, single turn
R19	5 kilohm linear potentiometer
SW1	7-station, spst, interlocking pushbutton switch
SW2	SPST toggle switch
T1	transformer, 12.6 Vct, 1.2A Radio Shack 273-1505
misc.	heat sink, 8-pin DIP socket, 2 binding posts, PC board, line cord, perforated, 1/4" standoffs, test leads, etc.

fig. 4. Rear view of ohmmeter.

construction

The power supply components are mounted on a 3-3/4 × 4-1/2 inch (9.53 × 11.4-cm) perfboard which is mounted on the meter studs. The PC board containing the measuring circuit components is mounted on the perfboard with two 1/4-inch standoff spacers. The range switch, mounted on the front panel, is a 7-station push-button type (I prefer this rather than a rotary type switch.) **Fig. 2** shows the foil side of the PC board and **fig. 3** shows the arrangement of parts on the component side of the board. **Fig. 4** shows the placement of all components as seen in the view of the rear of the ohmmeter while **fig. 5** shows the front view.

The meter scale is non-linear, as shown in **fig. 6**. At midscale the readings are equal to the standard resistors for the various ranges. To draw the scale a simple calculation can be made, knowing the meter's full-scale deflection in degrees. (Mine was 101 degrees.) The formula is:

$$\frac{B}{A+B} \times C = \textit{degrees deflection for } R_X = B$$

where A = standard resistor
B = unknown resistor
C = full-scale deflection in degrees

For example, using the × **1k** range $A = 10k$ and $C = 101°$, a 4k unknown will read:

$$\frac{4k}{10k + 4k} \times 101° = 28.86°$$

In building the prototype unit, I decided to mount the gain control on the front panel just in case some slight change in gain would be needed to correct for component value changes with time. So far I haven't found any reason to use it.

I chose to mount the ohmmeter on the back of my workbench in a central position from which four-foot long leads would reach test points where I could read the meter without parallax error.

calibration

Calibration is simple. Adjust the nulling pot for zero output of the voltage follower stage with the input terminals shorted. Then adjust the gain control for full-scale deflection.

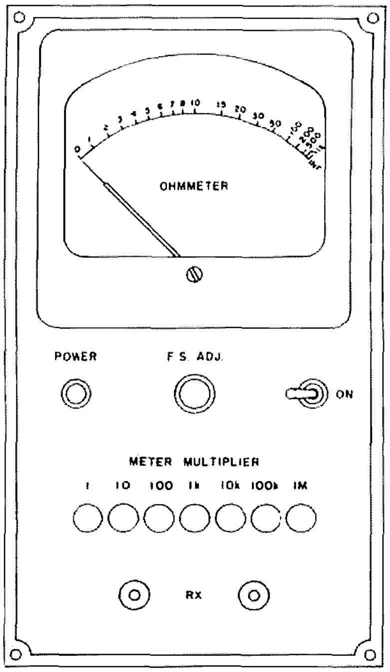


fig. 5. Front panel view of ohmmeter.

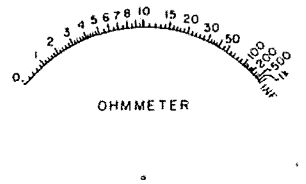


fig. 6. Typical meter scale layout.

operation

Switch on the power; the meter should deflect to full-scale. Connect the test leads across the unknown resistor, selecting the range which gives a reading below half-scale because the divisions are larger on the lower part of the meter dial.

If, on the ×1M range, jitter occurs at full-scale with the test leads not connected to an unknown, reverse the line plug. When using the ×1 range the resistance of the test leads introduces a small error which can be measured and subtracted from any fractional ohm measurement. Simply note the reading obtained with the test leads shorted at their tips. Then subtract this reading from the unknown resistor reading.

ham radio

ham radio TECHNIQUES

Bill Orr, W6SAI

antenna "neutralization"

As one cynic put it, "It's hard to be loud on 40 meters when you live on a city lot." *Many mini-antennas have* been designed for 40 meters (and for 80 meters, too) and some of them seem to work. However, they all pale into insignificance when measured against a full-size beam antenna.

So where does that leave the majority of hams who can't erect a big three-element 40-meter Yagi on a 150-foot tower? The only answer for them is to use a mini-antenna that *works.*

One knowledgeable Amateur who has wrestled with this problem is Les Moxon, G6XN,[1] who concluded that mini-beams constructed with close spacing (to conserve boom length) could never be optimized because the elements were overcoupled. His solution to this problem was to eliminate the excess coupling by means of a neutralization technique (**fig. 1**).

When Les was visiting in the San Francisco Bay area recently, I had the opportunity to speak with him about the technique. By the time he left, my head was abuzz with interesting ideas for mini-beam antennas for the low-frequency bands.

Alas, inertia and lack of free time prevented anything from being done. However, my good friend George Badger, W6TC, grasped the idea, combined it with some original thinking, and developed a mini-quad antenna for 40 meters that provides both good gain and excellent front-to-back ratio.

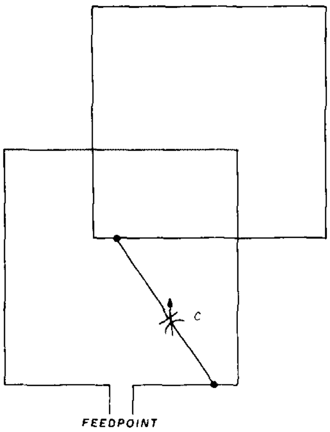


fig. 1. The G6XN neutralizing technique as applied to a 40-meter mini-quad. (Tuning and loading adjustments not shown.) The parasitic element in this case is a reflector. Note that the neutralizing capacitor is connected from one side of the parasitic element to the opposite side of the driven element. A small vacuum variable capacitor is used.

George had built a compact 40-meter quad that worked, after a fashion. Gain over a dipole was questionable, and the array didn't seem to have any front-to-back ratio. Bandwidth was very sharp, and the antenna operated over only a small segment of the 40-meter band before the SWR on the transmission line became too high for proper transmitter operation.

He solved the bandwidth problem with a simple tuning device for the quad loops, but the gain and front-to-back problems remained unsolved until G6XN proposed the antenna neutralization method.

Without going into construction details, the neutralization concept is shown in (**fig. 1**). It can be applied equally well to a close-spaced Yagi array (**fig. 2**). In essence, a capacitor is cross-connected from a driven element to a parasitic element and then adjusted for best front-to-back ratio of the antenna. Maximum forward gain and best front-to-back ratio seem to occur at the same setting of the capacitor. The tap point of the capacitor on the antenna elements seems to be less critical than the value of the capacitor. For W6TC's quad, the capacitance value is optimized for best performance.

Does the neutralization scheme work? W6TC's five-band DXCC proves that you *can* be loud on 40 meters when you live on a city lot! I'll provide more details on this interesting antenna development at a later date. Meanwhile, there's no

reason why you can't experiment with antenna neutralization on your mini-beam. I'd be interested in hearing from readers who try this unusual technique.

the coax jungle

Sometimes a little knowledge is worse than none at all. I've heard a lot of plausible stories about coax cables and I know that plenty of junk coax is being sold to unsuspecting buyers. But I didn't realize what a "jungle" exists with regard to the millions of miles of coax cable sold each year to hams, the military, industry, CBers, and others. The military, at least, usually knows what it gets for its money, as flexible coax cables are bought to a strict specification (MIL-17D or E). But how about the rest of us?

This all came to a head when I measured the loss of my coax transmission line running from station to antenna. According to all the published charts, my line loss at 14 MHz should be about 0.8 dB per hundred feet. My measurements, done with laboratory-type instruments, showed a loss of nearly 1.35 dB. What was the problem? Was the cable growing old? Or was it inferior cable to begin with? What should be the expected life of a coax cable? Five years? Twenty years? Forever?

a long phone call

Why not go directly to the coax cable manufacturer? One day, I did just that. I called one of the largest cable manufacturers in the United States and spoke at length to one of the production engineers who was an expert on coax cable manufacture.

I first asked about "military specification" cable. He said that if you want the best cable, that was the type to buy, but that it was also the most expensive because of the exhaustive tests the cable must pass before it gains military approval.

My unseen friend mentioned that there were perhaps a dozen large, top-notch coax cable manufacturers in the United States and about the

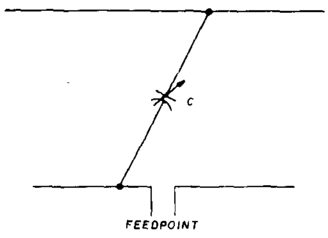


fig. 2. Top view of a compact two-element Yagi beam with neutralization capacitor. (Tuning and loading adjustments not shown.) Value of capacitor and tap points on the elements are adjusted for best front-to-back ratio after beam is adjusted for maximum gain.

same number of smaller, but good manufacturers. And finally, there were numerous "garage shop" operators who made up coax on demand to *any* specification. "After all," he said, "a coax cable machine isn't that expensive."

Part of the cost of the cable is the exhaustive tests and inspections demanded by the military; sometimes the cost of certification exceeds the cost of manufacturing. If you don't have to meet military specifications, the costs drop sharply...especially if you use second-hand cable machinery.

Consider RG-8/U cable, which is not in military use today. (Military gear is designed for 50-ohm termination, and the old style RG-8/U and newer RG-8A/U cables were 52 ohms.) "Of course, if you want," said the cable company spokesman, "we'll manufacture RG-8/U or RG-8A/U to a military specification, but we've had no call for that."

RG-8/U and RG-8A/U were replaced by RG-213/U (50 ohm) cable made to MIL-C-17D or E specifications. The specifications call for an inner conductor made of seven strands of bare copper wire, each strand 0.0296 inch in diameter. The dielectric is first-grade polyethylene, with no dirt or pinholes in it. One untinned copper braid jacket is used, providing greater than 90 percent coverage. He pointed out proudly that his particular non-military RG-8A/U had 97 percent coverage. The outer jacket was polyvinylchloride IIa, noncontaminating material. And the cable sold for well over 70 cents a foot in small quantities.

"What about coax that sells for about 30 cents a foot?" I asked. Well, that was a long story. Plenty of price-cutting is going on, but it is possible to manufacture an inferior grade of coax that looks pleasing to the casual observer. He had seen cheap cable with less than 60 percent braid coverage for sale. Moreover, the size of the braid wire was reduced from B.S. No. 36 to B.S. No. 36.5 or B.S. No. 37. ("You can buy wire *any* diameter you want," he emphasized.)

The next price-cutting technique is to use a less expensive jacket, such as used on the older RG-8/U cables. These jackets gradually contaminated the copper shield and he classified this old material as PVC-1. He'd seen cable like this one on the market.

Another cost-cutting step in making a cheap cable is to reduce the diameter and number of wires in the center conductor, and to also decrease the twist, or turns-per-inch, of the conductor. And, finally, a low grade of polyethylene insulation can be used. Some cheap cables use *all* these cost-cutting techinques.

During the manufacturing process, costs could be shaved if no attention were paid to the concentricity of the cable. Letting the center conductor "wander" a bit meant that less cable was rejected in final inspection and an older cable-making machine could be used for the job.

When the cable was completed, it might look like the *real* thing, but losses would be higher and cable life would be shorter than that of high-quality cable, and the impedance would probably wander well away from 50 ohms.

The engineer sighed. "I think much of the cable on the ham and CB market is sub-par quality; as long as price is the determinant, that's the way it's going to be."

To summarize, he rapidly summed up his observations for me:

1. *Be careful of RG8/U, RG-8A/U, or "RG-8/U type" cable. It has no controlling military specifications. RG-8A/U made by a reputable manufacturer — with its name and type number on it in addition to the RG nomenclature is probably OK...but it is almost as expensive to buy as the premium stuff.*

2. *Look up the specifications for the cable you intend to buy, and examine the cable before you buy it. Does it have the proper number of center strands? Is the inner dielectric uniform with no blemishes or spots? Does it have complete, or nearly-complete, braid coverage? Does it have a noncontaminating jacket? And most important of all, is it made by a known, reputable manufacturer? Be suspicious of underpriced cable because "there ain't no free lunch."*

3. *If you are starting out fresh, buy RG-213/U cable (50 ohms) which also has the manufacturer's name and type number on it instead of RG-8/U type cable (52 ohms). It may be nitpicking, but today's RG-213/U cable standard is 50 ohms. It is a military-approved, properly-manufactured cable. Again, make sure it is made by a reputable manufacturer. You can then be certain the cable is what it is claimed to be and know that it has been tested.*

long life for your coax

Once having bought good cable, how does the owner get maximum life from it?

1. *Keep the cable off the ground and make sure it can dry off after a rain. Because modern outer jackets are slightly hygroscopic, moisture can penetrate the jacket material, reach the outer braid, and cause corrosion.*

2. *Try to keep the cable out of direct sunlight; ultraviolet rays are damaging over time. For prolonged exposure to strong sunlight, the cable outer jacket should be a high molecular weight polyethylene with imbedded carbon black (expensive!).*

3. *Support the cable every ten feet or less. Don't let it sag on a long run.*

4. *Don't let the coax cable whip around in the wind. Repeated flexing is not conducive to long cable life.*

5. *Seal the ends of the cable. Use type-N (waterproof) fittings instead of the cheap, plentiful PL-259 plugs. Coat the terminations with non-acid type silicone rubber sealant. ("If it smells vinegary, that indicates acetic acid in the sealant. Don't use it.")*

6. *Don't step on the cable or otherwise flatten it. And don't bend it around a sharp radius. The minimum recommended bend radius is equal to ten times the outer diameter of the cable. (That's about a 5-inch radius for RG-8A/U and RG-213/U.)*

how long will it last?

I'd heard it said that coax cable should be replaced every few years, so I asked my friend if that was so. He replied that all cables deteriorate at greater or lesser rates depending upon use and abuse. Cable failure occurs when one or more specifications of the cable no longer meets the needs of the user. Sometimes the impedance of the cable changes; sometimes the cable loss becomes intolerable; sometimes the velocity of propagation changes to a degree.

I told the engineer that my RG-213/U, which was about 15 years old, had increased in loss by a measurable amount over the years and asked whether I should scrap it and buy new cable.

"That's up to you," he replied. "Coax cable doesn't fall apart all at once like the One Horse Shay. Deterioration is gradual. If you don't mind the increased loss, why spend the money to replace the cable? If you require every watt, you'd better get rid of it."

"Then this story of a five-year life, or a fifteen-year life for coax is unsubstantiated?" I asked. "That's right," he replied. He said he thought the story had begun because in the past, some branches of the military had routinely junked coax cable after it had been stored for a number of years. He didn't think that was true today.

"Well, if I were operating on 80 meters," he added, "I wouldn't worry too much about the grade of coax I had. Losses are relatively low at that frequency even on old, junky cable. At 20 meters, however, I would start to pay attention to cable loss. And at 2 meters, I would examine the coax I bought very carefully. I would measure the velocity of propagation of each section if I made matching transformers, and I would check cable loss every year or so. As you go higher in frequency, you can't be too careful!"

His parting shot was to remind me that large quantities of imported cable which have recognized U.S. markings but whose manufacturing specifications are unknown are now entering the market. "Have fun when you buy your next length of coax cable," he chided as we hung up.

free brochure

I have reprinted the EIMAC brochure covering the design of Pi and Pi-L networks for linear amplifier service. Write to me at EIMAC, 301 Industrial Way, San Carlos, California 94070 and ask for bulletin AS-30. (Two first-class stamps or two IRCs would be appreciated.) And for full information on the design and construction of linear amplifiers, I recommend the 22nd edition of the *Radio Handbook*, available through Ham Radio's Bookstore, Greenville, New Hampshire 03048 ($21.95 postpaid).

new EIMAC 3CX800A7 power tube

A free data sheet on the new EIMAC 3CX800A7 power tube is available from Varian EIMAC, 301 Industrial Way, San Carlos, California 94070.

reference

1. Les Moxon, G6XN, describes his antenna neutralization technique in his book, *HF Antennas for All Locations*, available from Ham Radio's Bookstore, Greenville, New Hampshire 03048 ($14 postpaid).

ham radio

EMI/RFI shielding: new techniques
part 1

Electronic pollution raises new interest in plastic shielding

This two-part primer on plastics discusses their use in suppressing the undesired effects of EMI/RFI (emitted electromagnetic interference and received radio frequency interference). Part 1 examines methods of EMI/RFI suppression; part 2 reviews the various methods of testing shielding effectiveness that employ RF techniques familiar to most Amateurs. Before delving into a discussion of methods of EMI/RFI suppression, let's see just what this interference is, what its adverse effects are, and why there seems to be a sudden surge of interest in this area despite what appeared to be little or no concern in years past.

Intense foreign competition has caused American manufacturers of electronic equipment to cut costs dramatically; as a result, metal cases have often been replaced with less expensive plastic cases. Plastics are basically insulators; consequently they provide virtually no EMI shielding unless they are specially treated.

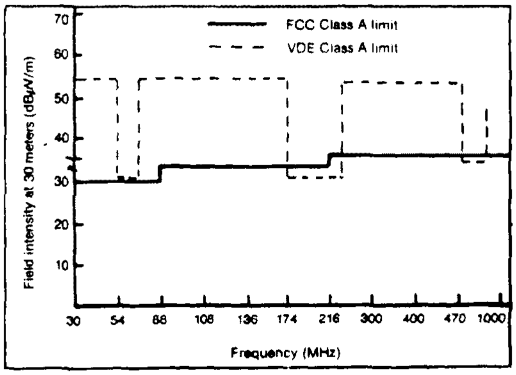


fig. 1. VDE vs. FCC standards for Class A products.

electronic pollution

Ten years ago there was only a small fraction of the number of electronic devices now available to consumers: electronic watches, calculators, home computers, video tape recorders and games, electronic toys, CB radios, microwave ovens, thermostats and home security devices, and a variety of controls on new cars. All of these devices emit electromagnetic radiation. In addition, the emphasis on portability in many of these items means that prolonged battery life through power reduction techniques — such as using switching voltage regulated rather than linear voltage regulated power supplies — has assumed far greater importance. Granted, the switchers are about twice as efficient as the linears, but because of their inherent mode of operation, i.e., switching current into an inductor, they are veritable EMI generators. Consequently, shielding becomes essential.

pollution effects

According to FCC chairman Mark Fowler, the number of electronic interference complaints now exceeds 80,000 per year.[1] For example, communications at an East Coast airport were hampered by interference from a "noisy" cash register in a drug store a mile away. In a western state, police communications at 42 MHz were disrupted by interference from coin-operated video arcade games. A home computer was found to affect its owner's TV as well as sets through-out the neighborhood. A fleet of new passenger buses *designed for urban* use were delivered, but during a test drive a driver tried to brake as

By Vaughn D. Martin, 114 Lost Meadows, Cibolo, Texas 78108

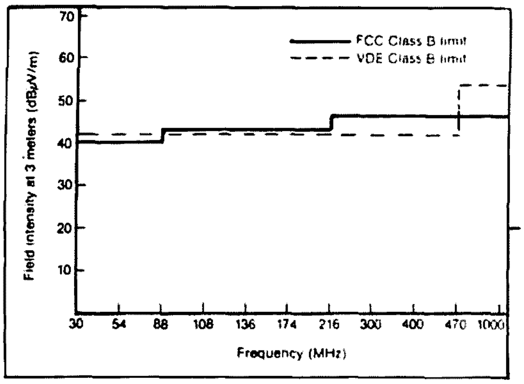


fig. 2. VDE vs. FCC standards for Class B electronic products such as computers and electronic games.

he went downhill and was astonished to discover he had no brakes until he reached bottom. Why? Because the microprocessor-controlled anti-skid devices controlling the brakes had been rendered useless by a commercial TV station's signal lobed in the direction of that particular hill. Once the electronic housings in the buses were shielded, their brakes worked.

background on FCC rule-making

On April 23, 1976, the FCC began a rulemaking proceeding (referred to as Docket 20780, Part 15, Subpart J) for setting realistic limits on VHF and UHF frequency emissions. The law included six categories: class, definitions, emission limits, verification, certification, labeling, and dates of compliance in dealing with radiated EMI/RFI in electronic and electrical industrial devices. In January 1979, a major computer manufacturer petitioned the FCC to relax some of these proposed standards. In response, the FCC relented on a few minor changes and rescheduled the July 1, 1980 compliance date. Other proposals presented are under consideration as well. While class A products (industrial electronic devices) are of interest, most concern seems to be focused on class B products: personal computers, electronic games and similar consumer items. (Certain inexpensive electronic games operating at frequencies less than 495 kHz are exempt from the new FCC ruling.)

Emission limits on class B devices are about three times as stringent as those placed on class A industrial products. Note in **figs. 1** and **2** how class A and B limits compare with the West German or generally accepted European standard, the VDE. The German post office has placed half-page ads in newspapers pointing out that items not meeting the specifications contained in the VDE regulation will not have a government proof number. Purchasers of devices without such numbers are liable to prosecution for electronic pollution; therefore, if an American product is to be marketed in West Germany it must be very EMI-proof. But the FCC *is* tightening up on emissions. For example, as of October 1, 1983, all manufacturers of class A and B products became subject to the FCC's request for random sampling of

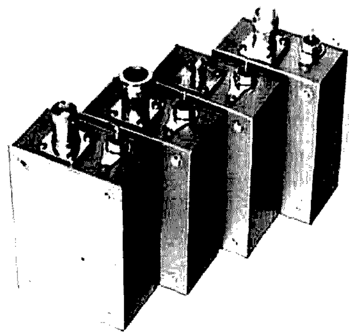


fig. 3. Typical small enclosures designed for EMI shielding.

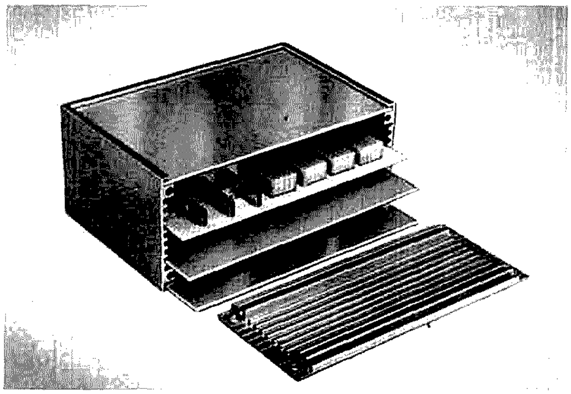


fig. 4. A metal case made with tightly fitting sides and end plates for EMI containment.

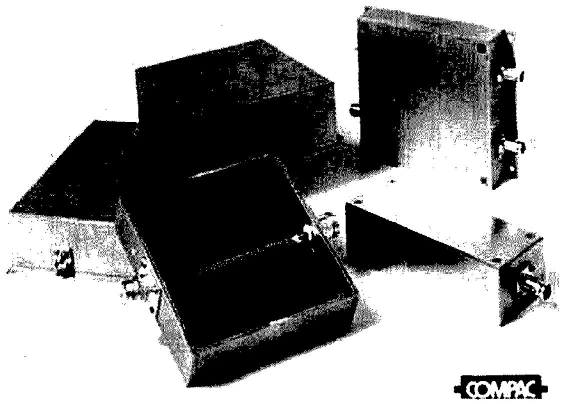


fig. 5. Nickel-plated EMI shielded cases.

table 1. *Methods of EMI shielding and/or containment.* (A) *lists the shielding effectiveness* (SE) *parameters of various* particle coatings, while (B) provides alternate methods of shielding.

(A):

coating*	coating thickness (mils)	sheet resistance ohms/sq.mm at 1 mil)	typical attenuation (dB)
plastic	125.0	0	0
aluminum/panel (for reference)	62.5	0	65-100
silver paint	1.0	0.01	65-70
silver/graphite (two coats)	0.2/1.0	0.01/100	54-77
copper	2.0	2.00	20-65
copper/graphite (two coats)	2.2	8.0/100	27-62
graphite	2.0	20.00	30-60
nickel	2.0	1.00	65

(B) alternate methods of shielding

1. Conductive elastometric materials
2. Conductive particle sprays and paints
3. Pressure sensitive adhesive backed tapes
4. Laminated heatsinks with paths for conducted EMI suppression
5. Wire mesh in strips and sheets
6. Zippertubing™ of coverings for cables that suppress conducted EMI and EMP (electromagnetic pulses)
7. *Conductive composite fillers* (*covered in Part 2 of this article*)

*Information supplied by PacTec Corp., Philadelphia, Pennsylvania

products to check compliance with established standards.*

the consequences of budget cutting

There is a twist to this story: a directive from FCC chairman Mark Fowler to Commissioner Anne Jones directed her to comply with the President's desire to cut their office's budget by 12 percent. In order to

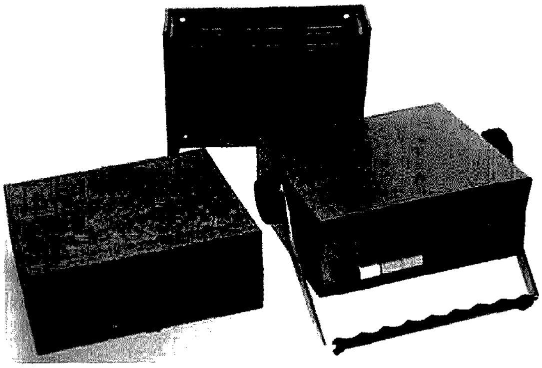


fig. 6. Plastic cases impregnated with conductive fillers to suppress EMI.

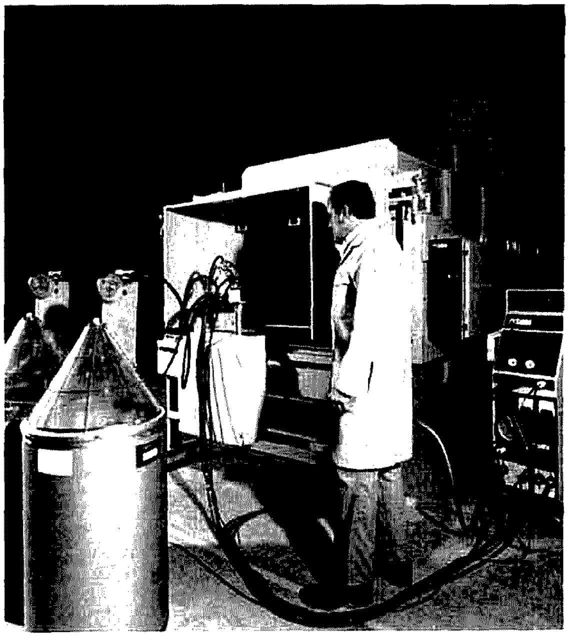


fig. 7. An X-Y axis programmable molten zinc sprayer.

*Any questions should be directed to your local FCC field office or to Art Wall, Office of Science and Technology, Federal Communications Commission, Washington, D.C. 20554.

comply with the directive, the FCC may decide to close its Laurel, Maryland, test lab. If this happens, will the FCC then contract out its work to an independent laboratory? And if it does, will total impartiality be guaranteed? With the compliance-auditing arm of the FCC eliminated, will manufacturers disregard the new FCC rulings and eliminate metallized plastic cases and power line filters in an effort to produce their products more cheaply?

shielding enclosures

In the absence of effective government regulation, what can Amateurs do on their own?

Surprisingly, the most cost-effective way to solve the electronic pollution problem isn't careful circuit design, but rather the use of well-shielded enclosures. While there are many ways to accomplish this (see **table 1**), all are essentially the same in that they are variations of treating enclosures. (The emissions previously referred to are naturally radiated emissions; conducted emissions are covered later, but suffice it to say for now that conducted emissions are best prevented by adequate filtering of power lines and other lines or cables leaving or entering the "box.")

Metal enclosures were the first to offer efficient EMI shielding. Today many companies still produce metal enclosures: Compac, for example, offers its RFT series, which includes built-in SMA, TNC, N, and BNC interchangeable RF connectors, (see **fig. 3**). Other manufacturers such as Vector Electric produce a series of aluminum Multi-Mod enclosures with tightly overlapping sides and end plates for EMI containment (see **fig. 4**). Like most metal cases, however, these are more expensive than plastic ones and generally offer less decorative, less protective outer finishes than plastic cases.

table 2. Companies specializing in EMI shielding.

	type products manufactured					
shielding manufacturers/ company addresses	metal cases	foil	mesh	applied coatings	magnetic devices	tubing for cables
Adhesives Research Inc. Glen Rock, Pennsylvania 717-235-4860	—	yes	—	—	—	—
Advance Process Supply Chicago, Illinois 312-829-1400	—	—	—	—	yes	—
Compac Deer Park, New York 516-667-3933	yes	—	—	—	—	—
Perfection Mica Co. Magnetic Shield Division Bensenville, Illinois 312-766-7800	—	—	—	—	yes	—
Stackpole Corp., Electronic Components Division St. Marys, Pennsylvania 814-781-1234	—	—	—	—	yes	—
Tafa Metallisation, Inc. Bow, New Hampshire 603-224-9585	—	—	—	yes	—	—
Technit Inc. Cranford, New Jersey 201-272-5500	—	—	yes	yes	—	—
Vector Electronics Co., Inc. Sylmar, California 213-365-9661	yes	—	—	—	—	—
Zippertubing Company Los Angeles, California 213-321-3901	—	yes	yes	—	yes	yes

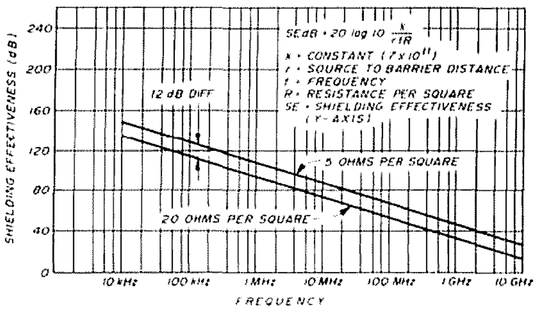


fig. 8. Shielding effectiveness vs. frequency-shielding effectiveness is proportional to the resistivity of thin film coatings, and increases with frequency.

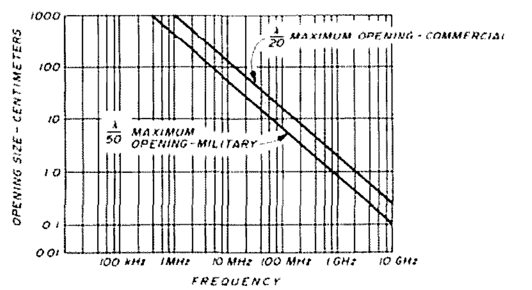


fig. 9. Hole size vs. radiated emissions frequency. Additional shielding materials are not required in a commercial enclosure if hole size is less than 1/20 wavelength, or 1/50 wavelength for military enclosures.

Fortunately, plastic cases can be treated to resist EMI as effectively as metal cases. This treatment process takes one or more of four forms: (1) the cases can be lined with a metal-plated or foil wire mesh screen (see **fig. 5**); (2) the cases can be painted or sprayed with conductive-particle plastic sprays; (3) the cases can be manufactured with conductive fillers impregnated in the plastic itself (see **fig. 6**); or (4), the cases can undergo a treatment process for the direct deposit of metals (see **fig. 7**).

The arc spray coating shown in **fig. 7** is a particularly interesting method used at Tafa Metallisation. Despite the fact that similar methods of using high temperature oxy-fuel plasma may damage some plastic surfaces, this method is so gentle that when molten zinc is applied to the surface of fresh fruit, the fruit will not be scorched. This seemingly impossible task is accomplished with a metal sprayer that looks very much like a paint sprayer but differs in that two small rods of metal stock are fed to the heat source and melted by an arc welding technique. This is accompanied by cool compressed air blowing fine atomized particles of molten metal onto the surface to be treated. This allows the plastic's surface to remain below 125° F (50° C), despite the fact that zinc itself requires a minimum temperature of 770° F (410° C) to even begin to melt. A 0.003 inch (0.0762

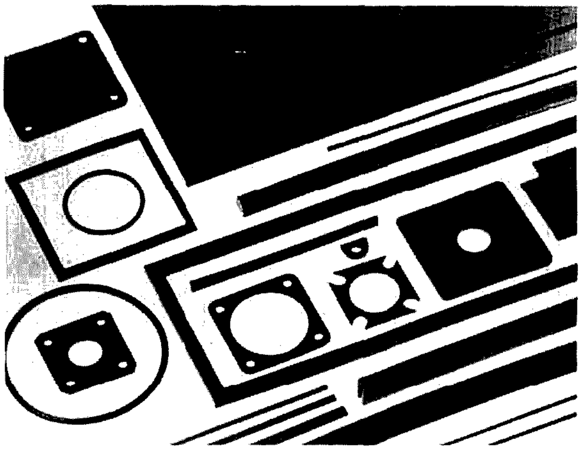


fig. 10A. Gaskets impregnated with conductive materials.

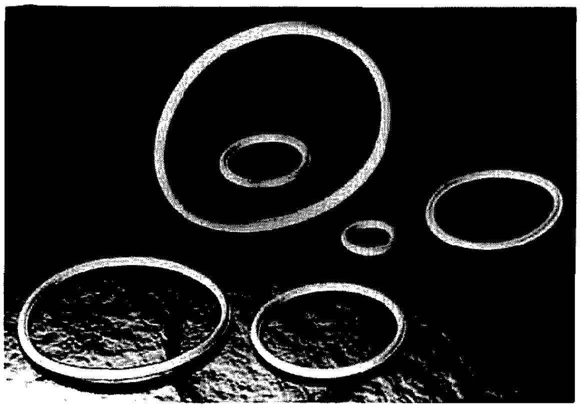


fig. 10B. Rubber grommets containing conductive particles.

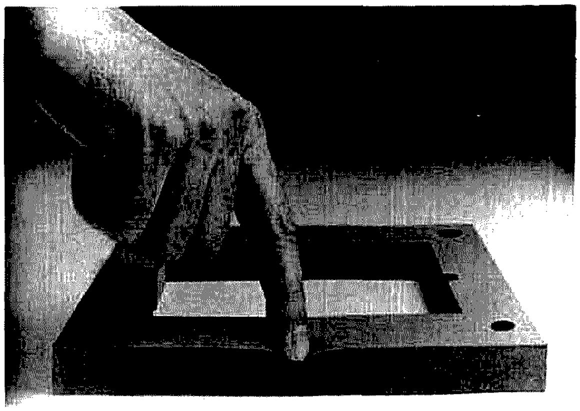


fig. 10C. A foam rubber cushion containing conductive particles.

mm) coating of zinc over a one-square foot surface requires about 3 ounces of zinc at an approximate cost of 19 cents. A 0.002 to 0.005-inch (0.0508 - 0.127 mm) — thick coating of zinc affords a resistivity sufficient for 50 to 90 dB of attenuation to radiation in the sub 100 MHz to 10 GHz range. This is important because EMI problems are essentially broadband in nature, so that shielding methods are devised for effectiveness against the highest potential frequency.

This arc spray coating is classified as either thin film (under 2.5 microns) or thick film (above 2.5 microns in thickness). Thin film coatings work particularly well at higher frequencies (see **fig. 8**) and provide shielding as a function of their resistivity. These resistive thin film coatings contain minute particles of silver, nickel, or carbon. Surface conductivity is therefore "adjustable" according to the type of spray selected, with mixtures using from 20 to 80 percent conductive materials to provide high conductivity. This is much more than commercial applications require (60 dB at 100 MHz). These coatings are quite fragile, though, and require an additional thin dielectric coating to avoid scratching or "fingerprinting" the freshly sprayed plastic surfaces. A scratch can be the "crack" in the box that permits EMI to escape.

specialized EMI problems

So far we have discussed the containment of EMI using various techniques, but have perhaps labored under the inaccurate assumption that any tight box will contain EMI. Generally, a tight box *will* contain EMI, but most pieces of equipment must have openings. These take the form of holes in the front panel through which the control shafts (potentiometers, variable tuning capacitors, and such) protrude. Additionally, LED or LCD displays or even a CRT often prove to be troublesome with their even bigger openings; larger pieces of equipment may have blowers, fans, vents, and louvers open to the outside environment. The cases themselves also often have discontinuities in seams or joints. Covers, gaskets, access ports, display windows, shaft holes, cables, and connectors all make it difficult to achieve optimum EMI/RFI integrity.

There is a relationship between their diameter and the frequency of radiated emissions (see **fig. 9**). The gaskets and rubber grommets that fit around these holes make the gap between the control shaft and the outside world much tighter (see **fig. 10**). In addition, these grommets are also often treated with conductive materials such as nickel, cobalt, iron, silver, or graphite. Many gaskets are made of wire mesh or conductive elastomers (supple, rubberlike plastic sheets — see **fig. 11**). Less supple metallic gaskets are therefore limited to straight-edge runs or openings with large radii. **Table 2** lists companies and products offered in this general area of EMI shielding. There are mechanical fittings made of wire mesh combined with silicone rubber or neoprene for cores that are both water- and gas-tight. There are also metallic tapes made with sticky or adhesive backs that can be applied to seams in an enclosure from which EMI might radiate. Glass that is relatively clear but has nonetheless been impregnated with conductive EMI constraining particles, (**fig. 12**), would be ideal in any application using a CRT.

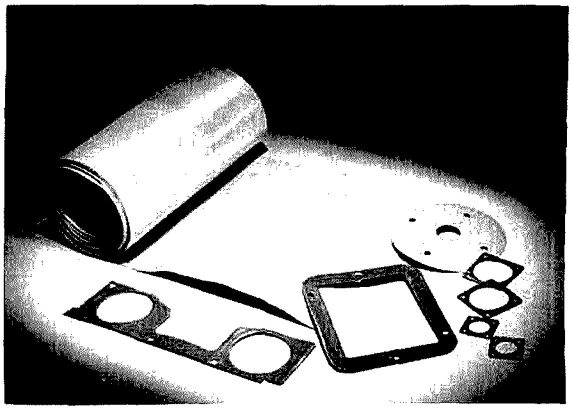


fig. 11. Supple elastomer sheets in roll form.

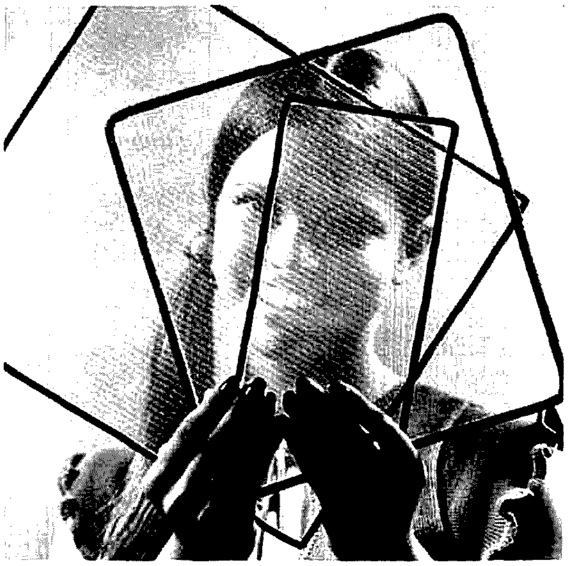


fig. 12. Glass impregnated with screens to contain EMI.

references

1. Dr. Theodore J. Cohen, N4XX, "CQ Interviews Mark S. Fowler, Chairman, FCC," *CQ*, March 1982, page 18.

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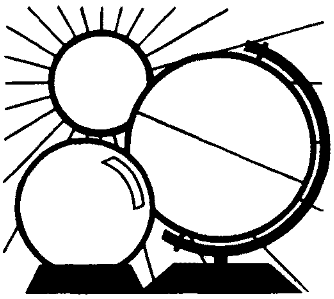

DX FORECASTER

Garth Stonehocker, K0RYW

last-minute forecast

Midwinter DX conditions are expected to be very good this first month of 1984. The higher frequency bands (10-30 meters) are expected to provide long skip and transequatorial openings on the 7th and possibly on the 20th. The lower bands (30-160 meters) are expected to provide excellent conditions throughout the month, with short skip during the daytime, between the above dates, and longer skip openings at night when the geomagnetic field is quiet. Geomagnetic ionospheric disturbances can be expected about the 1st, 10th, 18th, and 27th; the 5th, 13th, and 23rd are also possible disturbance dates, though with somewhat lower probability.

For EME and meteor burst communicators, the lunar perigee is on the 19th and a full moon appears on the 18th.

updating the foF2 formula

In the September, 1983, *DX Forecaster*, formulas were given for determining a mid-latitude local-noon foF2 (F2 layer critical frequency) from solar radio flux or sunspot numbers. The foF2 is the most variable of the propagation parameters that help define the maximum usable frequency (MUF) for a path between you and the DX QTH (MUF $\cong$ 2.5 foF2). Keep in mind that the MUF is about the optimum frequency for working that DX. These formulas are meant to be used for predicting "long term" (base) values of foF2 (for a few weeks or a month).

Changes in foF2 correlate with the daily solar flux changes over the previous three days. Delay in correlation (onset of foF2 change) depends on the magnitude of the flux change. Consequently these daily flux values can be used to update or "fine tune" the "long term" foF2 values discussed before. Furthermore, the rate of change of foF2 is only 30 percent the rate of change in solar flux and is used to alter the "long term (base) value."

but what about today?

To determine today's foF2, use your recorded values of the flux over the previous few days, noting how much it has varied. This variation, when multiplied by 0.3, equals the incremental increase (or decrease) in the long term value or base value of foF2. If you want a forecast of the foF2 that may be expected to occur several days ahead, use the last days' flux change. Small flux changes (up to 5 or 10 units) do not cause the ionosphere to change very much, so do not work too far back in time. The cause of the delay is the time the electrons take to drift or diffuse upward into the F2 region after being produced below 180 km by the solar flux (ultraviolet).

The flux — foF2 correlation is quite good in winter and equinoxal months. For lower latitudes the correlation is even better; for higher latitudes and summer months, correlation is poorer. The solar flux correlation with lowest usable frequency or absorption of the signal (signal strength) is *immediate*, with no delay experienced. Remember this correlation when using the lower frequencies in or near daytime.

band-by-band summary

Ten meters will occasionally be open, providing F2 long skip by the transequatorial one-long hop propagation mode (TEM). The openings will follow the sun during the day and into late evening. Geomagnetic disturbances will *enhance* this mode, as will high solar flux. Openings may favor southern Africa, South America, and Australia.

Fifteen meters can experience the same TEM modes as 10 meters with the openings being more frequent. World-wide DX is prevalent from after sunrise until well after sunset during the periods of high solar flux (listen to WWV at 18 minutes after the hour for reports on solar and geomagnetic conditions).

Twenty and thirty meters will be open most days and into the night to some areas of the globe with 1000-2500 mile long skip and some short-skip of 1200 miles near midday. Both propagation modes follow the sun across the sky; east, south, then west.

Thirty and forty meters are the transition bands, providing all-night propagation as well as some short-skip conditions during the daytime. Most areas of the world can be worked from darkness hours till just before sunrise. Hops shorten on these bands to about 2000 miles, but the number of hops can increase since signal absorption is low during the night.

Eighty meters offers ample opportunity for much DX work. Several stations have worked over 300 countries on this band. The band operates much like 40 meters except that the hop distances shorten to about 1500 miles. Noise from distant thunderstorms is low enough to make these bands a joy to work this time of year. The path direction follows darkness across the earth (east, south, then west).

One-sixty meters, similar to 80 meters, will provide multihop opportunities, though each hop shortens to 1000 miles.

ham radio

JANUARY

WESTERN USA

GMT	PST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	4:00	20	20	15	10	15	10	10	20
0100	5:00	20	20	15	10	15	10	10	20
0200	6:00	20	30	15	10	15	10	10	20
0300	7:00	20	30	15	10	15	10	10	20
0400	8:00	20	30	15	10	15	10	10	20
0500	9:00	20	*30*	20	10	15	10	10	20
0600	10:00	*20*	*30*	20	10	15	10	15	20
0700	11:00	*20*	*30*	20	15	15	15	15	20
0800	12:00	*20*	*30*	*20*	15	15	15	15	*30*
0900	1:00	*20*	*30*	*20*	15	*20*	15	20	*30*
1000	2:00	*30*	*40*	*20*	*20*	*20*	20	20	*30*
1100	3:00	*30*	*40*	*20*	*20*	*20*	20	*20*	*30*
1200	4:00	*30*	*40*	*20*	*20*	*20*	20	*20*	*40*
1300	5:00	*30*	*30*	*20*	*20*	*20*	*20*	*20*	*40*
1400	6:00	*30*	*30*	15	*20*	*20*	*20*	*20*	*40*
1500	7:00	*40*	*20*	15	*20*	*20*	*20*	*20*	*40*
1000	8:00	40	20	15	20	*15*	15	*20*	*40*
1700	9:00	40	20	15	20	*15*	15	15	*30*
1800	10:00	40	20	10	20	15	15	15	20
1900	11:00	40	20	10	15	15	15	15	20
2000	12:00	30	20	10	15	15	10	15	20
2100	1:00	30	20	10	15	15	10	10	20
2200	2:00	30	20	15	15	15	10	10	20
2300	3:00	30	30	15	15	15	10	10	20

MID USA

MST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN	CST
5:00	30	20	15	10	15	10	10	20	6:00
6:00	30	20	15	10	15	10	10	20	7:00
7:00	30	30	15	10	15	10	10	20	8:00
8:00	30	30	15	10	15	10	10	20	9:00
9:00	30	30	20	15	15	10	10	20	10:00
10:00	30	*30*	20	15	15	10	15	20	11:00
11:00	*30*	*30*	20	15	15	15*	15	20	12:00
12:00	*30*	*30*	20	15	15	15	15	*30*	1:00
1:00	*30*	*30*	20	*20*	*20*	15	20	*30*	2:00
2:00	*30*	*40*	*20*	*20*	*20*	*20*	*20*	*30*	3:00
3:00	*20*	*40*	*20*	*20*	*20*	*20*	*20*	*30*	4:00
4:00	*20*	*40*	*20*	*20*	*20*	*20*	*20*	*30*	5:00
5:00	*20*	*30*	*20*	*20*	*20*	*20*	*20*	*40*	6:00
6:00	*20*	*30*	15	*20*	*20*	*20*	*20*	*40*	7:00
7:00	*20*	*30*	15	15	*20*	*20*	*20*	*40*	8:00
8:00	*20*	*20*	15	15	*20*	*20*	*20*	*40*	9:00
9:00	30	20	15	15	*15*	15	*20*	*40*	10:00
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1:00	30	20	10	10	15	10	15	20	2:00
2:00	30	20	10	10	15	10	10	20	3:00
3:00	40	20	15	10	15	10	10	20	4:00
4:00	40	30	15	10	15	10	10	20	5:00

EASTERN USA

EST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE CARIBBEAN S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
7:00	30	20	15	15	15	10	10	20
8:00	30	30	15	15	15	10	10	20
9:00	30	30	15	15	15	10	10	20
10:00	30	30	15	15	15	10	10	20
11:00	30	30	20	20	15	10	15	20
12:00	*30*	*30*	20	20	15	10	15	20
1:00	*30*	*30*	20	20	15	15	*15*	20
2:00	*30*	*30*	20	20	*20*	15	*20*	*30*
3:00	*30*	*40*	20	20	*20*	20	*20*	*30*
4:00	*30*	*40*	*20*	20	*20*	*20*	*20*	*30*
5:00	*30*	*40*	*20*	*20*	*20*	*20*	*20*	*30*
6:00	*20*	*40*	*20*	*20*	*20*	*20*	*20*	*30*
7:00	*20*	*30*	*15*	*20*	*20*	*20*	*20*	*40*
6:00	*20*	*30*	*15*	*20*	*20*	*20*	*20*	*40*
9:00	*20*	*30*	15	15	*20*	*20*	*20*	*40*
10:00	20	*20*	15	20	*20*	15	*20*	*40*
11:00	20	20	10	20*	*15*	15	*20*	*40*
12:00	20	20	10	15	*15*	15	15	*30*
1:00	20	20	10	15	15	15	15	*30*
2:00	20	20	10	15	15	10	15	20
3:00	20	20	10	10	15	10	10	20
4:00	30	20	10	10	15	10	10	20
5:00	30	20	15	10	15	10	10	20
6:00	40	20	15	10	15	10	10	20

The italicized numbers signify the bands to try during the transition and early *morning* hours, while the standard type provides the MUF during 'normal' hours.
*Look at next **higher** band for possible openings.

the weekender

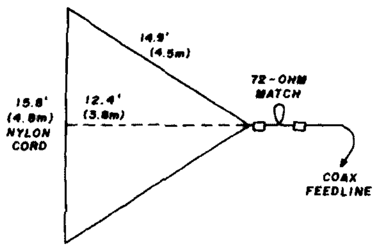

fig. 1. V antenna for 2 meters.

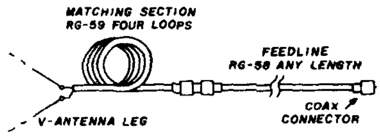

fig. 2. V antenna feedline details.

the 2-meter V-antenna

I am always on the lookout for new antennas for HF and VHF applications because they represent projects that can be assembled on a weekend at minimum cost, with maximum enjoyment. Recently, the V antenna came to my attention as a possible 2-meter portable radiator.

The V antenna (see **fig. 1**) consists of two long wires spreading apart from one end in an angle of less than 90 degrees at the apex (known as the apex angle). The maximum direction of radiation bisects the angle, requiring careful placement for best results. The leg length determines the overall gain and directivity of the system as well as the apex angle. I chose a leg length of 2.25 wavelengths and a 65-degree apex angle resulting in a 15-foot long wire antenna for 2 meters. (Charts are available if you wish to use a different leg length.)[1]

To determine the length in feet for a frequency of 146.9 MHz, I used the following expression:

$$L = 984\,(N - 0.025)/F$$

where L = length in feet
N = number of wavelengths
F = frequency (MHz)

Solving for:

$$L = 984(2.25 - 0.025)/146.9$$
$$L = 14.9\,\text{feet} = 14\,\text{feet}\ 11\,\text{inches}$$

construction

I used No. 22 gauge stranded insulated hookup wire for the antenna for several reasons. The wire is very flexible and does not end up in a tangle during storage. More importantly, the wire is available at any electronic supply store and is not expensive.

By Thomas M. Hart, AD1B, 32 Westwood Terrace, Westwood, Massachusetts 02026

A recent article by Bill Orr[2] discussed the problem of feeding V antennas. I found his solution quite satisfactory for 2 meters. A 50-ohm coax line services the system, matched by a 1/4-wavelength section of 75-ohm coax.

The length can be found using the following:

$$L = 246(V)/F$$

where V = velocity factor
F = frequency (MHz)
L = length (feet)

$$L = 246\,(0.66)/146.9$$
$$= 1\,\text{foot}\ 1\,\text{inch for RG-59 (73 ohm)}$$

The Orr article recommends forming the matching section into four loops to "decouple the outside of the line from the antenna currents."

The completed antenna, run as a horizontal V, loaded well (SWR 1.1:1) and enabled me to easily access the desired repeaters. So, if you are interested in a different solution to the perennial antenna problem, try this conversation piece. In terms of efficiency, the reference texts indicate that a V with 2.25 wavelength legs has a theoretical gain of 4.5 dB over a halfwave dipole. The V antenna will provide a substantially better signal than the usual 1/4-wave coat-hanger vertical.

references

1. *The Radio Amateur's Handbook*, ARRL, 1980, pages 20-7 to 20-9.
2. William I. Orr, W6SAI, "Ham Radio Techniques," *ham radio*, July, 1983, page 42.

ham radio

the ham notebook

three circuits for repeaters

Unmanned repeater operation requires the use of remote control circuits. **Figs. 1, 2,** and **3** illustrate three popular circuits that enhance repeater operation: the autopatch phone line interface, the phone auto answer and ring indicator, and the TouchTone™ sequence decoder. All three are constructed of easy-to-find components.

autopatch phone line interface

Fig. 1 is a schematic diagram of a repeater-to-phone line interface for use as an autopatch. Its function is to provide for the receiver-to-phone line and phone line-to-transmitter link, with both using an opamp for gain. Q1 and Q2 control which audio path is active; when the receiver squelch sense line goes above 1 volt, Q1 conducts. This enables the receiver audio to pass the phone line. At the same time, Q2 turns off and the phone-to-transmitter audio is disabled. The receiver squelch input, therefore, must provide a high (above 1 volt) when an input is present. When the receiver squelch sense input goes below 0.5 volts, Q1 turns off and Q2 conducts, passing phone-to-transmitter audio and disabling the receiver audio path.

R1 adjusts the receiver-to-phone line audio and R2 adjusts the phone-to-transmitter audio level. R3 is an optional adjustment to allow for passing phone audio to a TouchTone™ decoder.

Transformer T1 isolates the electronics from the phone line and thereby maintains a balanced phone line. Relay RL1 acts as the on/off hook interface for the patch. When RL1 is energized, the phone line will be seized and audio can be passed to and from the phone line to the repeater receiver/transmitter. A low on the "line seize" will energize RL1. Make sure that diode CR2 is in place as shown; otherwise damage to the circuit that energizes RL1 could occur when the relay is turned off.

Diode CR3 and the 100k resistor force the audio circuits to pass receiver-to-phone audio and disable the phone-to-transmitter audio when the autopatch is not in use. This prevents any audio which might be produced from the autopatch circuit from going to the repeater transmitter. Several different types of op amps can be used. In this case, LF353 is used. In choosing a different op amp, the primary consideration should be one of stability.

This circuit has been used by many repeater groups with much success. Problems which might be encountered are primarily with dialing phone numbers. This autopatch circuit will provide for a very flat audio response with low noise and distortion. However, with the typical HT or mobile rig, it is often found that the rolloff characteristics will produce imbalance in the high to low group levels in the TouchTone™ audio. The phone company is becoming more and more

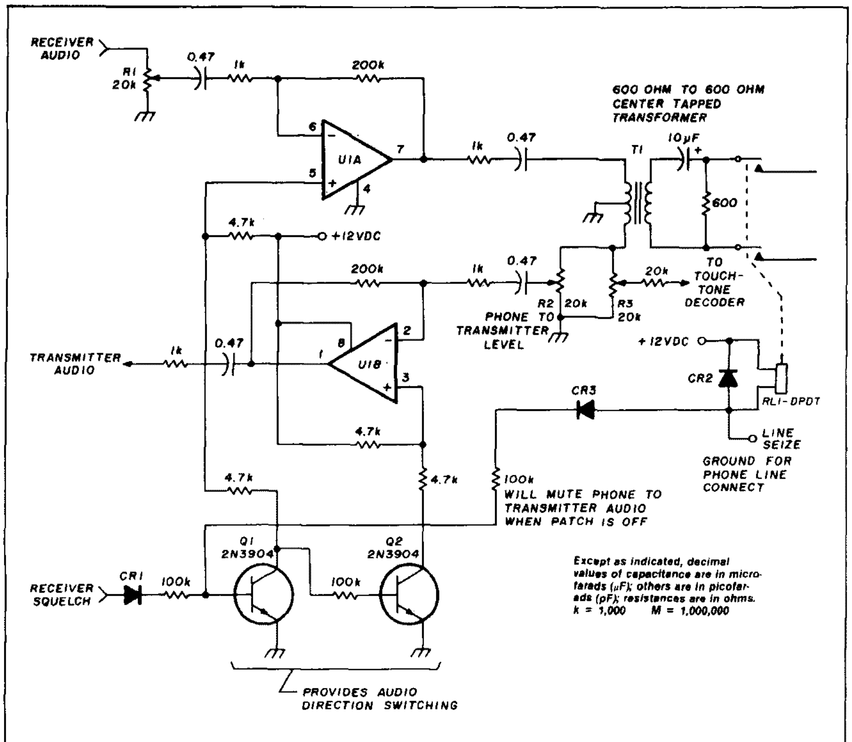

fig. 1. Autopatch (repeater-to-phone line) interface schematic diagram.

sensitive about this and is using equipment which is fussier about what it will accept than previous equipment. Thus, even though one may have plenty of level, the tones will often be rejected by phone company equipment. I assume this is done to provide for a greater degree of falsing prevention of their TouchTone™ decoders. The solution is solved by providing equalization that emphasizes the high response with about a 6 dB per octave rolloff. That is to say, a two-to-one ratio between the 697 Hz tone (lowest low group TouchTone™) and the 1209 Hz tone (lowest high group TouchTone™).

If you need an autopatch — and most repeater users do — this is a good and reliable circuit.

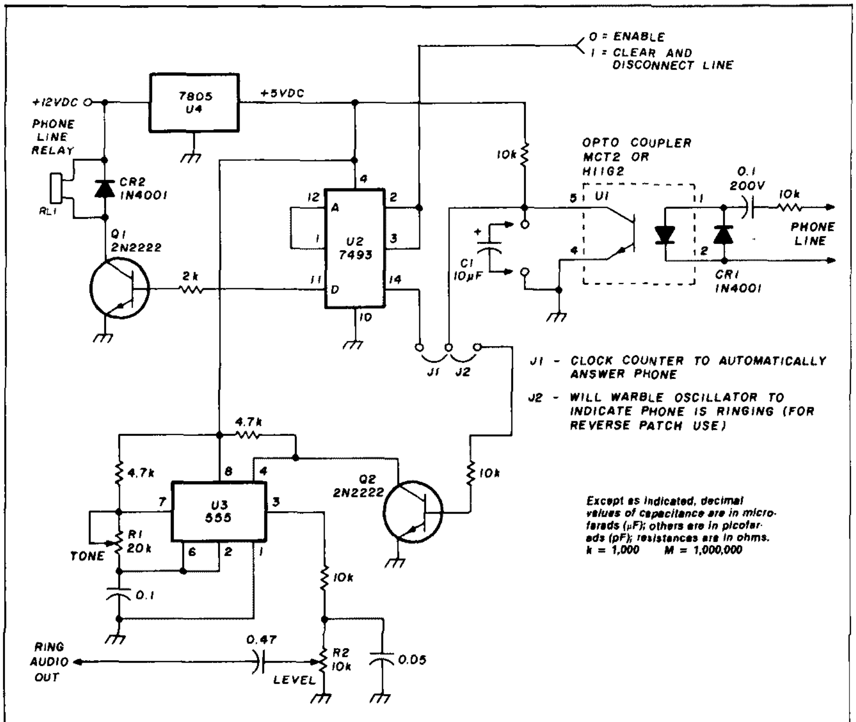

fig. 2. Auto answer or ring indicator schematic diagram.

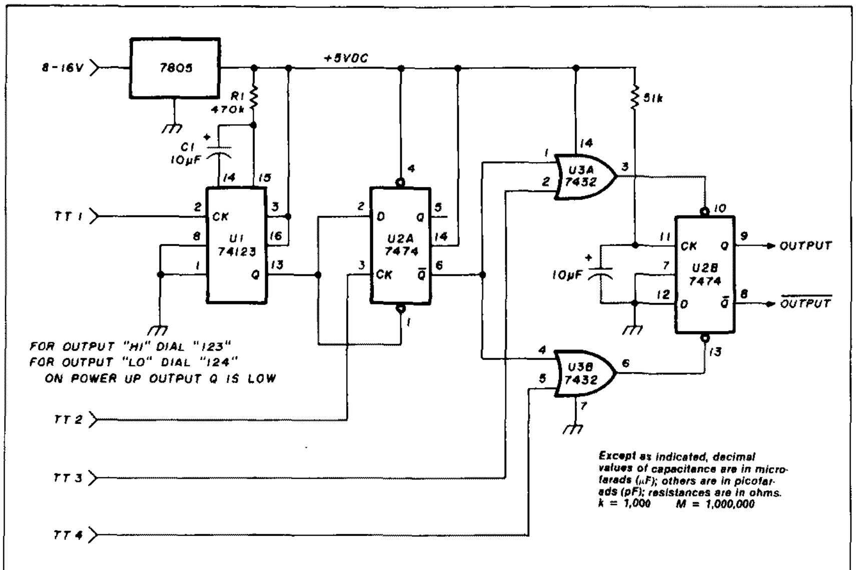

fig. 3. TouchTone™ sequence decoder schematic diagram.

phone auto answer and ring indicator

Fig. 2 is a schematic diagram of a ring detect circuit for automatic phone answering or tone generation for reverse autopatch use.

U1 is an opto-isolator consisting of an LED and phototransistor. When the LED lights, the phototransistor conducts. The diode is connected across the phone line through a 10K resistor and a 0.1 µF capacitor. This combination is transparent to the phone line, which is a must as far as the phone company is concerned. When the phone starts ringing, the ring signal produces pulses which force U1 to conduct and generate pulses in turn. This output is coupled to U2 and the pulses are counted. When the count reaches 8, U2's output goes to a logic high and forces Q1 to conduct closing the line seize relay, RL1.

The output of U1 is also coupled to U3 which is a 555 timer IC operating in the oscillator mode. When pin 4 goes high the oscillator will produce a tone at a frequency determined by R1. R2 is the level adjustment for this tone. This tone can be used to signal a repeater listener when the phone is ringing for use in a reverse autopatch. In the reverse patch mode, jumper J1 should not be connected in order to prevent U2 from answering the phone, stopping the ringing of the phone.

Jumper J1 determines whether the phone is answered automatically and J2 forces the tone generator to produce a tone on each ring of the phone. If capacitor C1 is not used, the output pulses from U1 will be at the phone ring rate (20 or 30 Hz depending on the phone system). This will force the phone to be answered

almost immediately and the tone generator to give a warbling tone. However, if C1 is inserted, a single pulse will be given for each ring. Thus, the phone will be answered on the eighth ring and the tone generated will be constant while the ring is taking place.

The "clear" input to U2 clears the ring counter when a logic high is present on this line. This input should be connected to the decoder logic, enabling a caller or repeater user to force the disconnect of the phone when desired.

TouchTone™ sequence decoder

Fig. 3 is a schematic diagram of a TouchTone™ sequence decoder circuit. Its purpose is to take active low inputs from a TouchTone™ decoder and react to a proper sequence of digits. The proper sequence is determined by which TouchTone™ digits the user connects to the sequence decoder inputs TT1, TT2, TT3, and TT4. The function of this circuit is most useful for repeater builders that require some sort of logic control such as enabling and disabling functions on the repeater.

The circuit operates on the principle of requiring the digits to arrive in a proper sequence, preventing a function from being activated should a random sequence be entered. For example, if TT1 = TouchTone™ digit 1, TT2 = digit 2, TT3 = digit 3 and TT4 = digit 4, then the code for forcing the output to a logic high is *123*. To force the output low the code is *124*. If the sequence is dialed in any other order, the sequence decoder will not react to it (for example, dialing *213* will do nothing).

U1 is a "one-shot" which reacts to a pulse from the TT1 input by forcing its Q output, pin 13, to a high for a time determined by R1 and C1. Larger values keep high longer. Using the given values Q remains high for 4 seconds. If TT1 is strobed again during this period the 4-second period will restart. However, if TT1 is not strobed again, Q will revert back to a logic low. When the Q output is in the low state, no sequence decoding can take place. U1 acts as both a timer and a first digit seeker. After the first digit is dialed the remaining digits of the code must be dialed within 4 seconds. U2A decodes the second digit of the sequence. Upon dialing the second digit, the Q ($\overline{Q}$ not) goes low, enabling the TT3 and TT4 inputs of U3. If TT3 is dialed next, pin 3 of U3 will go low, producing a latched output on U2Bs Q output, pin 9. If TT4 is dialed as the third digit, the Q output will be forced low.

Pin 9 of U2B is a latched output and will remain in its state as long as power is maintained or until the proper sequence is dialed forcing Q to another state. Of course if an "ON" code is dialed and the state of the decoder is ON, no change will take place.

C2 and R2 force the Q output of U2B to a low on power up. This feature is most desirable in cases where power may be lost and no one is around to reset the circuit to the desired state, which is the case in most unmanned repeater sites. However, if the desired output state is a high on power up, the user should use the $\overline{Q}$ output. In this case the ON/OFF codes are reversed from the above description.

As shown, the ON/OFF codes are three digits each. However, if only a two-digit code is desired, removing the A part of U2 and connecting U1's $\overline{Q}$ pin 4 output to pins 1 and 4 of U3 will produce a two-digit decoder with the ON code being TT1 and TT3.

The OFF code becomes TT1 and TT4. If more than three digit codes are desired, all one must do is add D-type flip-flops after U2A and before U3 connecting U2As Q (not $\overline{Q}$) to the added flip-flops D input and its $\overline{Q}$ output to U3. For each added flip-flop the code is increased by one digit.

The suggested digits for setting up the codes is determined primarily by how the decoder is to be used. In the case where TouchTone™ might be used for other functions such as dialing numbers on an autopatch, it is most desirable to make sure that a telephone number will not produce the same sequence of digits as a controlling code. Thus, if a controlling code were *123* and a phone number of 844-1235 is dialed, the sequence decoder would react. This can be prevented by making sure that at least one of the digits in a code is not a 0 through 9. Thus, one should use a # or a *, or A, B, C, or D, if 16 digit TouchTone™ pads are desired, for at least one digit of a code.

Ron Wright, N9EE

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Coming Events

ACTIVITIES

"Places to go..."

South Bend, INDIANA: Hamfest Swap & Shop, January 8, first Sunday after New Year's Day at Century Center downtown on U.S. 33 Oneway North between St. Joseph Bank Building and river. Industrial history Museum in same building. Carpeted hal acre room. Tables $3 each. Four lane highways to door from al directions. Talk in freq: 52-52, 99-39, 93-33, 78-18, 69-09, 145.43 145.29.

MICHIGAN: The Cherryland Amateur Radio Club's 10th annua Swap and Shop, February 11, 8 AM to 1 PM, Immaculate Con ception School Gym, Traverse City. Register at door. Talk in on 146.25/85. For information call K8YVU, Jerry Cermak at (616 947-4848.

NEW YORK: Yonkers Electronics Auction sponsored by th Yonkers ARC, Sunday, January 22, 9 AM to 3 PM, Lempko Hall 556 Yonkers Avenue. Admission $3.00 each. Children under free. Auction starts 10 AM. New and used equipment. Club com mission on successful sales 10% on first $100, 5% on remainde Unlimited free coffee all day.

OHIO: Mansfield Mid-Winter Hamfest/Auction, Sunday, Febru ary 12, Richland County Fairgrounds, Mansfield. Doors open t public 8 AM. Tickets $2.00 advance; $3.00 at door. Tables $5.0 advance and $6.00 at door. Half tables available. Talk in o 146.34/94. For information, tickets/tables SASE to Dean Wrass KB8MG, 1094 Beal Road, Mansfield, Ohio 44905. (41 589-2415.

OHIO: Cincinnati ARRL '84 State Convention and Flea Marke February 25 and 26. Registration $5. Flea market $4 per spac both days. Forums, meetings, vendors, Wouff Hong, banque Hospitality suite Friday and Saturday nights. Write: Cincinna ARRL '84, POB11300, Cincinnati, OH 45211 or call (51 825-8234.

VIRGINIA: Frostfest '84, Winter Amateur Radio and Comput Show, Sunday, January 15, Better Living Building, Virginia Sta Fairgrounds, Richmond, 8 AM to 4 PM. General admissic $4.00. Flea market space $3.00, tables $3.50 additional. Boot for commercial exhibitors. Contact N4DDM (804) 272-8206 Richmond Frostfest, Box 1070, Richmond, Virginia 23208.

OPERATING EVENTS

"Things to do..."

JANUARY 8-9: Rats Nest and Crooked Stick IV, an antenna experimenter's sprint contest. 2100Z January 8 to 0100Z January 9. Frequencies: CW 21.060 to 21.200 MHz. SSB 21.350 to 21.450 MHz. Rats Nest & Crooked Stick Antenna: 100 ft. max. of single conductor wire. Feedline will not count if it is coaxial cable. Antenna is limited to 20 ft. high. Transmitter Power: 250 watts or less (DC input). Exchange: Name, QTH, type of antenna, I.A.R.C. member or not. Contest entries must be submitted by February 1, 1984. For more information: SASE to Issaquah ARC, Bob Farnworth, KB7NV, 6822 - 131st Avenue S.E., Bellevue, WA 98006

FEBRUARY 2-8: The Michigan Technological University ARC and the Copper Country Radio Amateur Association announces a radio celebration of our Winter Carnival festivities in the northernmost part of Michigan's Upper Peninsula. A certificate will be issued to all Amateurs who make contact with any participating ham in the Copper Country between 0000 February 2 through 0000 February 8. Frequencies: RTTY - 3.630, 7090, 14.095. CW 3.705, 7.085, 14.085, 21.085, 21.185. Phone - 3.930, 7.285, 14.305, 21.385, 28.685. On CW listen for CQ Winter Carnival. Send your QSL with 3-20¢ stamps for p&h to: Howard Junkin, N8FHF, 106 W. South Avenue, Houghton, MI 49931

FEBRUARY 4-5: Vermont QSO Party sponsored by the Central Vermont ARC (W1BD), 2100Z Feb. 4 to 0700Z Feb. 5 and 1100Z to 2400Z Feb. 5. Frequencies: Phone - 3910, 7230, 14260, 14320, 21360, 28570, 50110, 144.2. CW - 3530, 3730, 7030, 7130, 14060, 21060, 21160, 28060. RTTY - 3620 & 090 other RTTY sub-bands. Exchange: VT stations send QSO number, 2 letter county designator. Others send QSO number and state or province. Scoring: VT 1 point per phone contact, 2 points per CW or RTTY. Times states plus provinces plus ARRL countries. Others 1 point per phone contact, 2 points per CW or RTTY times number of VT counties. *A station may be worked 3 times per* band, once each on phone, CW or RTTY. Separate awards to Vermont and non-Vermont stations. Send logs/facsimiles, name, address, Vermont county NLT March 1, 1984 to: D. Nevin, KK1U, W. Hill, Northfield, VT 05663.

FEBRUARY 4-6: New Hampshire QSO Party, sponsored by the NH ARA, 1900Z Feb. 4 to 0700Z Feb. 5 and 1400Z Feb. 5 to 0200Z Feb. 6. Exchange signal report and QTH. Suggested frequencies: *Phone 3.935, 3.975, 7.235, 14.280, 21.380, 28.575,* 50.115, 145.015. CW 1.810, 3.555, 3.730, 7.055, 7.130, 14.055, 21.055, 21.130, 28.055, 28.130. RTTY 3.625, 7.085, 14.085, 21.085, 28.085. Logs must be postmarked by March 15, 1984. Include large SASE for results. Mail to: Pete Cantara, KI1M, 19 Haverhill Street, Hudson, NH 03051.

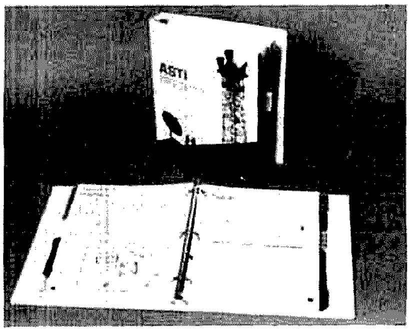
book and product

REVIEWS

ASTI

TI, Terrestrial Interference, is one of the most important problems facing users of the satellite service. Commercial and home TVRO owners have found that because of the relative newness of the technology they are using, there is little available in the way of information on terrestrial interference and how to avoid it.

Recognizing TI as a major problem, the folks at Microwave Filter set about the task of putting together a how-to book that would help identify and define problems and then set about the task of eliminating them. This book — *The Avoidance/Suppression Approach to Eliminating Terrestrial Interference of TVRO Earth Stations*, by Glyn Bostick, Tom Fannetti and William Johnson, is based upon the actual experience of the authors in their efforts to reduce the problems caused by TI. The authors also obtained a significant amount of input from the fledgling TVRO industry through symposiums and technical meetings held around the country.



During the early days of the TVRO industry, the common solution for a TI problem was to change the location of the system. But, now due to the explosive growth of the industry and high costs, this is not always a practical solution to the TI problem. Because of siting or budgetary limitations, it is often necessary to locate systems in less than optimal sites. Thus was born the ASTI project. The answers on how to reduce or eliminate TI and maximize system performance are found in ASTI.

Chapter 1 lays the ground work for TI avoidance and provides a brief summary of the rest of the book. Chapter 2 provides the reader with an illustrated summary of TVRO operation as it applies to the distribution of TV programming. Chapter 3 is a soup-to-nuts description of how TVRO works, starting with the uplink, transponder operation, and TVRO installation. This is most important because a complete understanding of the system is necessary to recognize how and where susceptibility will come. With that information in mind, chapter 4 details and describes potential TI sources by function and frequency. Chapter 5 deals in the symptoms of TI. Plenty of photos are provided so the reader can relate his unknown problem to known and quantified problems.

Chapters 6 through 11 deal with the actual site selection, antenna/LNA and other components in the system. Complete coverage is given to ensuring that the system is engineered properly, the first time through, to reduce TI problems to an absolute minimum. Some of the methods of TI suppression discussed are a bit severe, such as pits, fences, and other forms of artificial microwave barriers. However, some are absolutely necessary to eliminate the really thorny TI problems. The authors have gone to a great deal of trouble to discuss and eliminate some misconceptions that exist about TI shielding. In some cases, microwave shields are more expensive than they should be because of a lack of solid technical information. The authors give a complete description of the optical principles that control microwave radiation. And upon this framework, how innovative solutions can be derived to solve TI problems. Chapters 12 and 13 deal with how to eliminate unavoidable TI; Chapter 12 discusses what filters are available, their ability to counter TI, and their application at critical points in the system. Chapter 13 deals with "worst case" situations and takes full advantage of the authors' practical field experience in TI reduction.

Finally, chapter 14 deals with SMATV techniques for satellite and master TV systems. Techniques are outlined for avoiding and suppressing the interference that often comes from this kind of system hybridization.

This book is an absolute *must* for the professional TVRO and satellite earth station community. I would also recommend ASTI to home TVRO owners as an invaluable resource book. Even for those who do not now have a TI problem, construction of TI-producing systems continues daily. A trouble-free system today could become an unusable system tomorrow. The retail price for *ASTI*, $125.00, may sound a bit high but is quite reasonable in light of the wealth of information found within the book. (Besides, when you've spent more than $4,000 to install a TVRO system, the price of *ASTI* seems inexpensive, compared to a microwave technician's time in troubleshooting a TI-plagued system. MFC also provides an ASTI update service for $60/year that will keep you fully informed on all the latest TI problems and solutions.

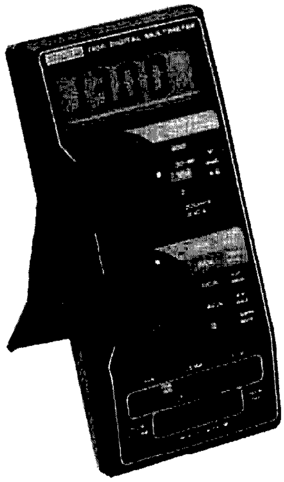
For more information on *ASTI*, contact MFC at 6743 Kinne Street, East Syracuse, New York 13057.

N1ACH

Circle #301 on Reader Service Card.

Keithley Model 130-A digital multimeter

The latest addition to Keithley's product line, the Model 130-A Digitial Multimeter, reached my desk the other day for review. I have been looking over quite a few multimeters over the last several months for *ham radio* and I must admit that upon opening the box, I found the Model 130-A to be the most aesthetically pleasing that I have seen. Functionwise, color has



little to do with performance. Ergonomically, however, a well designed, color-coordinated piece of test equipment will eliminate eye strain and fatigue that can reduce a test technician's overall performance.

Another interesting ergonomic feature is that the unit is designed to comfortably fit into your hand. Should the need arise to change scales or units to be measured, when the unit is in your left hand, it can be done easily and conveniently with the left thumb. This allows the user to hold the probes in position while the scales are being changed. Keithley also provides a handy carrying case with a belt loop. The small size means that you can stow the Model 130-A into your tool box without displacing too many tools.

The display for the Model 130-A uses an easy-to-read, 3/5 inch VCD readout. The large readout combined with work stand makes the 130-A digital multimeter right at home in your workshop, as well as being an invaluable tool for field service.

Keithley also has a full line of accessories for the Model 130-A that includes a temperature probe that will measure from −55 to 150 degrees C, high voltage probe that is rated up to 40 kV, a 50 ampere shunt, a clamp-on AC current probe, and of particular interest to Radio Amateurs and other RF people an RF voltage probe for measuring voltages from 100 kHz to 250 MHz.

The Model 130-A is rated at a maximum common mode voltage of 500 volts and is de-

signed and operated from 0 to 50 degrees C. It *is powered by a single 9 volt battery which is* rated at 100 hours usage with carbon zinc cells and 200 hours usage with an alkaline cell. The unit is 7.1 × 3.1 × 1.5 inches (178 × 78 × 38 mm) and weighs a mere 10 ounces (283 grams). It is priced at $145 and comes with battery, test leads, and instructions.

specifications

DC volts	200 mV to 1000 V (±0.25% + 1d)
AC volts	200 mV to 750 V (±1% + 3d) 45 - 500 Hz
Resistance	100 n ohms to 20 M ohms (±Avg. 275%)
DC amps	2 mA to 10 A (±Avg. 75%)
AC amps	2 mA to 10 A (±Avg. 2.4%)

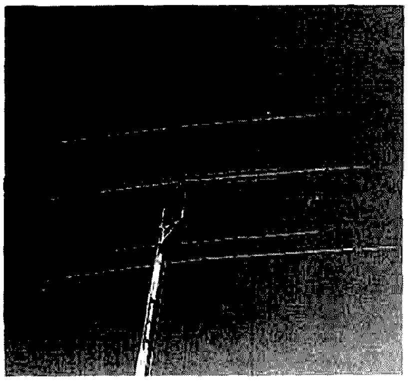
This is an excellent product and will be right at home in either your ham shack or on-the-job. For more information, contact Keithley Instruments, 28775 Aurora Road, Cleveland, Ohio 44139.

Circle #302 on Reader Service Card.

de KA1JWF

Explorer 14

The Explorer 14 is a three-band (10, 15, and 20 meters), four-element (two parasitics), Yagi antenna that provides complete band coverage *without* retuning. Telex/Hy-Gain, the manufacturer, was able to achieve this performance by incorporating a multi-section driven element and separate reflectors for 10 and 15/20 meters. (The single director uses 15



and 20 meter traps). It is relatively light (45 pounds, 20.4 kg), can be installed by one person (see **installation**), does not require any length/spacing readjustments after initial preset and is fed directly by 50-ohm coaxial cable — there's no gamma match to adjust. Its small size (14-foot boom) is an attractive feature for those who must be concerned with neighborhood aesthetics, while mechanically it accounts for only 7.5 square feet of wind surface area. Important specifications include a maximum gain of 8.8, 8.0, and 7.5 dBi for 10, 15, and 20 meters, respectively; a worst-case *VSWR of less than 2.0:1 from band edge to* band edge; and an average F/B ratio of 16 dB on 10/20 meters and 19 dB on 15 meters. It handles maximum legal power and is at DC ground potential including coaxial inner conductor.

para-sleeve system

The driven element consists of three sections: a center element and a closely spaced front and rear sleeve. The para-sleeve system (patent pending), basically an open-sleeve dipole, uses a trapped 15 and 20 meter driven element and *two* parallel elements for 10 meters. A Hy-Gain balun provides the balanced-to-unbalanced transformation while maintaining DC continuity between the antenna and cable inner and outer conductors.

construction

I enjoy assembling kits that are complete, have a well-written and diagrammed manual, and have been engineered for trouble-free assembly. In designing this kit, the staff at Hy-Gain must have carefully thought through each step, anticipated problem areas, and corrected for them as needed. Construction is simple; I have to admit that my two children (ages 7 and 9) built most of the antenna — and were usually one step ahead of me. All parts fit as described and were of good quality (stainless steel for the brackets and hardware).

Of course, the usual common-sense techniques are called for. Read through the manual carefully; clear a space in which to work; read the manual; unpack, count, and identify parts; read the manual, and proceed. You might have noticed that I've said "read the manual" three times; Hy-Gain recommends this as a good number. (I have personally seen "experts" assemble antennas incorrectly because they "knew" what to do.)

At one point in construction I was ready to pick up the phone and call Hy-Gain for two missing aluminum sections until I discovered that the manufacturer conserves packing space by telescoping some sections within others. In one particular case, the ID/OD dimensions were so close as to make the inner piece "disappear."

There are quite a few repetitive steps in constructing this antenna that lend themselves to production line procedures: compression clamp assembly, element-to-boom bracket assembly, etc. This is where my young workers excelled. The manufacturer suggests that at least five hours be allowed for assembly, and though we weren't in a race, I believe we beat that figure by at least half an hour. One of my children remarked, "Don't worry, Dad, we'll have you on the air by tonight." (Visions of climbing towers by moonlight had overtaken me.)

Final assembly is simpler if you drive a temporary 5-foot length of mast into the ground and attach the boom and boom-to-mast brackets to it. This permits eye-level installation of

parts and easy alignment of elements after assembly. This, naturally, is the best time to check and double-check all dimensions, (not when it's 100 feet in the air and 20° below zero).

I felt that there was a little too much play in the sleeve spacer clamping arrangement due to the use of 5/8-inch clamps — the next size smaller might have been better. Also, since no ·eadjustment of element lengths and spacings s ever called for, assembly would have pro- ceeded even faster if the manufacturer had narked the boom and color-coded the element sections. But then again this additional step is easily done, and by leaving it to the assembler, the manufacturer probably keeps cost down.

installation

I had remarked earlier that the antenna was ·elatively light. To prove it, I mounted the Ex- plorer 14 to a fifty-foot push-up TV mast. It withstood high winds and remained intact until disassembled several months later. I don't ·ecommend TV mast installation to the timid — t takes some resolve and a bit of muscle. Mine settled at the 42-foot level with only a few feet overlap of mast sections and four sets of polypropylene guys. The manufacturer recom- nends that it be mounted at least 30 feet above ·oofs or metallic structures if a low VSWR is to be achieved. I agree completely. Normal tower nounting procedures and precautions are imply described in the manual.

performance

The antenna test "range" at K2RR in Milford, New Hampshire, consisted of at least one completely open acre with no other anten- nas or structures (metal or otherwise) closer han 1½ wavelengths at the lowest operating requency — 14.0 MHz. The Explorer 14 was nounted at 42 feet above a low conductivity ground (New Hampshire isn't called the Gran- te State for nothing). The antenna used for A-B comparison exhibited at least 1 dBd gain oward Europe at a fairly low angle (estimate between 20 and 30 degree takeoff angle). A sim- ple coaxial relay allowed for rapid comparison.

Initial tests involved verification of the nanufacturer's VSWR specifications for the ntire three bands. The test setup included a ird wattmeter and a SWR bridge (used separ- tely). At the input to a 100-foot run of new 52- hm coaxial cable (Belden 8214), a VSWR of .0:1 was not exceeded at any frequency on 0, 15, or 20 meters and at many points was elow 1.5:1.

Both the gain and front-to-back evaluations vere qualitative, unfortunately, but in all for- vard gain tests the received signal strength vas greater from the Explorer 14 than from the eference antenna when both were aligned in he preferred direction. Since a bi-directional eference antenna was used, a rapid qualitative ont-to-back evaluation was possible. Signals vere weaker from off the back of the beam han from the standard antenna, as was ex- ected. To provide accurate gain and F/B figures, a calibrated test setup and better knowledge of the reference antenna and site are required.

30 and 40-meter add-on

The Explorer 14 can be operated on 30 or 40 meters as well with the addition of the Hy-Gain QK-710 conversion kit. The add-on kit includes a 20-meter trap, additional tubing, stainless steel hardware, and another well-illustrated and written manual. Depending on which of the two kits you choose, the driven element assembly resonates on either 30 or 40 meters (besides 10, 15, and 20 meters). It adds 3.5 or 6 pounds, to the overall weight of the Explorer 14 using the 30 or 40 meter kit, respectively. It provides rotary dipole performance with the advantage of a single coaxial feed still only re- quired for *four* band operation. (The add-on kit was not available at time of testing.)

field testing

In "field testing" the Explorer 14, I estab- lished a one-month goal to contact at least 100 countries during *non*-contest periods. It act- ually took only three weeks (June 18 - July 9)

to accomplish this. An abbreviated list of some of the "catches" is listed below. (A complete list is available; *send* SASE).

5H3SG	5Z4PR	UI80AA
UF6FER	UK8BAA(UH8)	UL7VBA
UK5OAA(U0)	YI1BGD	AP2ZA
C30LAB	A92F	VU2GI
UD6BR	SV1NA	3B8DB
J28DN	XT2AW	UJ8AP
A4XHZ	TU2AZ	

conclusion

The Explorer 14, model number EX-14, provides low VSWR broadband Yagi performance on 10, 15, and 20 meters. It allows the constructor to build it once, on the ground, without need of laborious readjustments for different mode operation. It is light enough for one-man installation, though an additional helping hand or two would be highly welcomed. Upon dismantling, no hardware showed any signs of corrosion — all the more surprising considering the acidic pH of New Hampshire rain. If you enjoy putting kits together and want a good performing no-tune Yagi, this definitely is one way to go.

For further information contact Telex/Hy-Gain, 9600 Aldrich Avenue, South, Minneapolis, Minnesota 55420.

K2RR

new Hamtronics® catalog

The new 1984 Hamtronics mail order catalog of supplies for VHF/UHF/OSCAR enthusiasts and two-way radio shops is now available. This 36-page, two-color catalog features many new products, including an expanded line of FM repeaters and accessories such as power amplifiers, DTMF tone decoder/controllers, and autopatches. Also included are the popular lines of FM and AM receivers, FM transmitters, VHF and UHF transmitting and receiving converters, space shuttle receivers, 800 MHz scanner converters, preamps, and other products.

For a free copy, contact Hamtronics, Inc., 65F Moul Road, Hilton, New York 14468. (For overseas mailing, please send $2.00 or 4 IRCs.)

Circle #303 on Reader Service Card.

technician's tool kit

The JTK-86 "Premier," new from Jensen Tools, is a full-service technician's tool kit in a compact zipper case. It includes a selection of tools rarely found in zipper case kits, such as a complete 13-piece 1/4 inch drive socket set with ratchet, spinner handle and extensions; a Vise-Grip locking plier; a 9-blade fold-up hex key set, and an 11-blade feeler gauge.

The JTK-86 is available in a vinyl or leather case with handles or in a case of rugged Cordura nylon with three roomy outside pockets for meters, test leads, or service manuals. Two pockets measure 5-1/2 × 9 × 2-1/4 inches, and the third measures 10-1/2 × 9 × 2-1/4 inches.

The JTK-86 "Premier" Jensen's finest zipper case kit, is available with or without a Fluke 8021 Digital Meter or a Triplett 310 VOM.

For more information or a free catalog of other Jensen kits and cases, write Jensen Tools, Inc., 7815 S. 46th Street, Phoenix, Arizona 85040.

Circle #305 on Reader Service Card.

720-channel handheld receiver

FDK International Corporation has begun to market what is said to be the world's first PLL-synthesized 720 channel hand-held AM airband receivers, the ATC-720/SP series. Designed for air traffic control and Amateur purposes, the series ATC-720/SP series employs PLL-synthesized circuitry for accurate frequency selection of 720 channels between 118-136 MHz in 25 kHz steps. The light weight (11 ounces; 315 gr.) and small size (6-5/8 × 2-1/4 × 1-3/4 inches; 169 × 58 × 43 mm) of the ATC-720/SP allows the user maximum portability in operating fields. Supplied with flex rubber antenna, Ni-Cd battery pack, and AC-charger, the ATC-720/SP features an adjustable squelch level to eliminate background noise on the AM mode. Low battery consumption allows 6 hours of continuous operation. A BNC aerial connector DC-charger and shoulder case are available as optional accessories.

For information, contact FDK International Corporation, 10-2, Kaji-cho 2-chome, Chiyoda-ku, Tokyo 101, Japan.

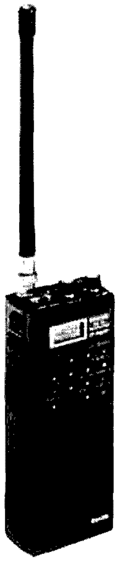
Circle #304 on Reader Service Card.

new handhelds

The new IC-02A and the IC-02AT two-meter handhelds are now available from ICOM. These compact multifeatured handhelds are the same compact size as the IC-2A series, but have features found on no other Amateur handheld.

The IC-02A and the IC-02AT are designed to be compatible with all existing IC-2A accessories plus new accessories that will make them unique. An important feature of the IC-02A series is that it features *32 PL* tones built into the unit as *standard*. These tones are programmable from the front panel pad, and may be used with any frequency selected, or may be stored in memory and recalled along with the frequency at any time.

Any frequency on 5 kHz spacing in the 2-meter ham band may be called up in the IC-02A. All frequency entries as well as control functions for memory, scanning, etc., are selected by the 16-button pad on the face of the radio. Included are priority watch, scanning of both memories and programmable band scan, and DTMF on the IC-02AT model. The unit features ten memories which store frequency, PL tone, offset and offset direction, and an internal lithium battery backup. The priority channel is a unique feature to the IC-02A and IC-02AT, as well as the custom LCD readout with an S-meter function, unique to the ham industry.



The IC-02A series will run at 3 watts with the standard BP3 battery pack, or at 5 watts with an optional high power battery pack. A long-life battery, 8.4 volts at 800 mA, will be available to double the working time of the standard 3-watt output unit. Batteries may be charged a variety of ways.

The IC-02A series has an environmentally sealed case with "O" ring seals to protect it

NEW products

against dirt and moisture. A heavy-duty aluminum back provides heatsinking for the 2 watts of power. A power connector is supplied on the top of the unit. Twelve volts applied there will power the unit as well as charge the battery pack.

For further information, contact ICOM America, Inc., 2112 116th Avenue, N.E., Bellevue, Washington 98004.

Circle #306 on Reader Service Card.

WARC bands kit for Yaesu FT-101

A new kit which provides receive/transmit capability on all three WARC bands for all models of the FT-101 except the ZD is now available from Fox Tango Corporation. While only the 10 MHz band has been authorized for use to date, little additional effort or expense will be needed to add all the bands while the circuit changes for 10 MHz are being made. In addition to making the old 101 ready when the new bands become available, the added capability increases the trade-in value of the set. Based on a tried and tested design by G3LLL, the WARC bands kit is complete with all needed crystals, relay, switch, and detailed instructions for moderately easy installation.

For further information, contact the Fox Tango Corporation, Box 15944H, West Palm Beach, Florida 33416.

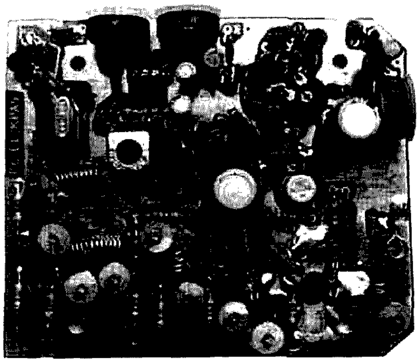
Circle #307 on Reader Service Card.

mobile ATV module

P.C. Electronics has released a 1-watt AM UHF ATV transmitter module board (Model KP5A) which, when mounted in an aluminum box atop a portable color camera, allows the camera operator to move freely. Coverage of up to a mile is typical under most conditions; 50 miles has been done from an airplane. Applications include video from radio-controlled model airplanes or robots, computer video, base station remotes, weather radar video, or any application in which cables would be impractical.

The KP5A is a wired and tested board capable of full color and sound. It comes standard with one crystal on either 439.25 (east) 434.0 (west), or 426.25 MHz. Its power requirement

is 13.8 VDC at 280 mA. The board size is only 4 × 3.25 inches (10.16 × 8.26 cm); its price is $159. Additional crystals are $15 each. Buyers must hold an Amateur license of Technician class or higher.

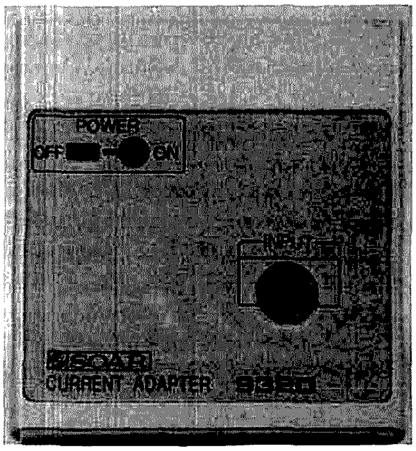


For further information, contact P.C. Electronics, 2522 Paxson Lane, Arcadia, California 91006.

Circle #308 on Reader Service Card.

three new adapters

Three new adapters designed to work with all DMM's having a 10 megohm or better DC input resistance and a 200 mV DC range are available from North American Soar. *Model 9320*, shown above, is a combination AC and DC current adapter complete with a clamp-



around Hall effect sensor. It can measure AC current to 150 amperes and DC current to 200 amperes without breaking into the line; it's priced at $79.00, with battery.

Model 9310 is a temperature adapter priced at $79.00, battery included. Switch-settable to read degrees in centigrade or Fahrenheit, it will work with all "K" type bimetal sensor probes.

NEW products

Model 9330 is a capacitance measuring adapter with two sets of input connectors, banana jacks for large or in-circuit devices, and pin-insert jacks for direct device capacitor tests. Priced at $44.00 with battery, this device can measure capacitance from 2 nF through 200 μF on six ranges.

For more information, contact North American Soar Corporation, 1126 Cornell Avenue, Cherry Hill, New Jersey 08002.

Circle #309 on Reader Service Card.

solar systems

Photowatt International, Inc., a leading manufacturer of "total" solar electric systems, offers a laminated solar panel module rated at 55 watts, 16.5 volts at 3.4 amps that features 5-inch silicon cells. Other modules ranging in power from 7 watts to 35 watts are available. The Solar Eclipse Series modules have a power output of 7, 12, and 25 watts and are laminated in a durable bronzed frame with charcoal gray Tedlar backing. Modules are wired for either 6 or 12-volt applications.

Photowatt also manufactures a photovoltaic regulator used for system control. These power control units are the only regulators that offer current and voltage meters, low-battery indicator, lightning protection, and additional components to protect the PV systems. The PCU's are in NEMA 3R, rain-tight, lockable enclosures. Ground and pole mount support frames are available for one to ten modules per structure. Batteries, wires, hardware, J-boxes, accessories and installation instructions are also supplied. Customized photovoltaic systems are available for immediate shipment. Computer sizing quotations are available upon request to match customer's specific PV power requirements.

For details, contact Photowatt International, Inc., 2414 West 14th Street, Tempe, Arizona 85281.

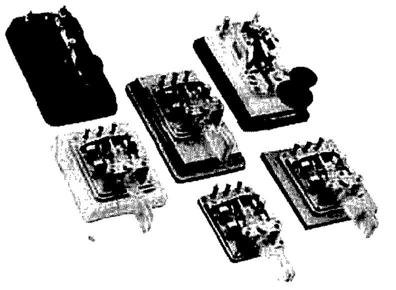
Circle #310 on Reader Service Card.

hand keys and keyer paddles

Guild, one of the most respected names in musical instruments and related electronics, recently became the sole distributor for the Hi-Mound line of iambic keyer paddles and hand keys.

Hi-Mound paddles feature silvered contacts with full spacing and tension adjustments on all

models. Three of the iambic paddles have *heavy*, slip-resistant bases (one of solid marble), while the fourth is a paddle assembly which can be mounted on the base of your choice or built into an existing keyer. The hand keys, in addition to retaining the classic look, also have silvered contacts and a unique tension adjusting system.



A clear plastic dust cover is included with each key, and all are protected by Hi-Mound's *one-year* warranty, backed by Guild.

For more information, contact Guild at 225 West Grand Street, Elizabeth, New Jersey 07202.

Circle #311 on Reader Service Card.

SSB/FM microphones

A *new* cardioid unidirectional microphone especially built for SSB and FM communications is available from Heil, Ltd.

Two models are available. The HM-5 features a die-cast metal base and heavy-duty flexible goose-neck stand. The MM-5 is a comfortable hand-held unit for use in mobile or fixed operations.

Both feature Heil's new HC-3 cartridge and offer enhanced intelligibility and articulation and clean, natural reproduction of voice transmissions. Sensitivity is measured to −70 dB; response, 300-4000 Hz; impedance, 2000 ohms.

For more information, contact Heil, Ltd., Marissa, Illinois 62257.

Circle #312 on Reader Service Card.

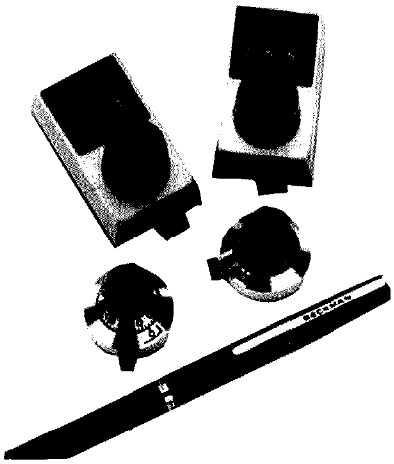
count on these

Two new series of turns-counting dials for panel control of precision potentiometers are now available from Beckman Electronic Technologies Group (BET). Ten new analog dials, series 2650 and 2660 Duodial® are available in the popular 1/8", 1/4", and 6 mm shaft sizes in either a black housing with white numerals or in an anodized case with black numerals. The 7/8" dials count up to 15 turns.

The eight new digitals, series 2100 and 2200 Digidial® are available in either the popular 1/4" or 6 mm shaft size and are available with a clear or black anodized housing. They count up to ten turns and are available in eight three- or four-digit models.

NEW products

Both series feature a positive locking mecha nism that prevents accidental changes due t shock or vibration. They mount directly on th potentiometer shaft to prevent backlash an are designed to compliment the Beckman Heli pot® potentiometer.



For complete information, contact Beckmai Instruments, Inc., Beckman Technologie Group, 2500 Harbor Boulevard, Fullerton, Cali fornia 92634.

Circle #313 on Reader Service Card.

new cable

Nemal Electronics has just introduced a ne cable for both the home and commercial sate lite TV market. This new cable eliminates a the separate control and RF cables and bundle them together into a single easy-to-handle covered package. Nemal's new SCC-921859 cable is covered with a high-strength, non-cor taminating vinyl outer jacket suitable for eithe direct burial or laying on the ground. Insid there is one Mil-Spec RG-59 RF cable with a 9 percent copper braided shield. There are fiv No. 22 copper strand wires and two No. 2 wires that are shielded with drain for sensitiv signal activator circuits. Finally, two heavy-du ty No. 18 wires are included for motor contro By eliminating the number of separate wires the chances of accidental system damage ar reduced. Installation time is also reduce because of the ease and convenience of run ning one cable rather than several. The price i tentatively set at 75 cents per foot and $495.0 per thousand feet.

For more information, contact Nemal Elec tronics, 12240 NE 14th Avenue, North Miami Florida 33161.

Circle #314 on Reader Service Card.

new MOS for 150 MHz power amplifiers

Motorola has introduced an extensive new line of RF Power MOSFETs offering output power in the range of 5 watts to 125 watts at 150 MHz. Manufactured with TMOS technology, these devices are designed for 28 volts operation in VHF amplifiers, with a range of 9 to 13 dB minimum gain. The devices are designated MRF134, MRF136, MRF171, MRF172, and MRF174.

These new TMOS devices exhibit higher gain, high input impedance, lower noise figure and lower intermodulation distortion than bipolars. This helps improve the performance of amplifying circuits and, in some cases, lowers the overall cost by requiring fewer stages of amplification.

specifications

device	output power (watts)	minimum gain (dB)
MRF134	5	11
MRF136	15	13
MRF171	45	12
MRF172	80	10
MRF174	125	9

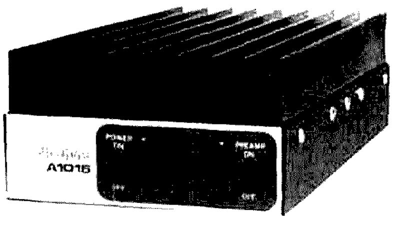
For more information, contact Motorola Semiconductor Products, Inc., P.O. Box 20912, Phoenix, Arizona 85036.

Circle #315 on Reader Service Card.

6-meter amplifier

Mirage Communications has recently been granted FCC type acceptance on their new 6-meter amplifier, the A1015. The solid-state unit is rated at 150 watts out for 10 watts in and comes with a list of features that includes a built-in receiver pre-amp, "all mode operation" (FM, SSB, and CW) and antenna changeover capabilities.

Designed to fill the growing demand for a 6-meter amplifier compatible with a variety of different 10-watt radios now in use, the unit sells for $279.95.



For more information, contact Everett Gracey, Mirage Communications, P.O. Box 1393, Gilroy, California 95020.

Circle #316 on Reader Service Card.

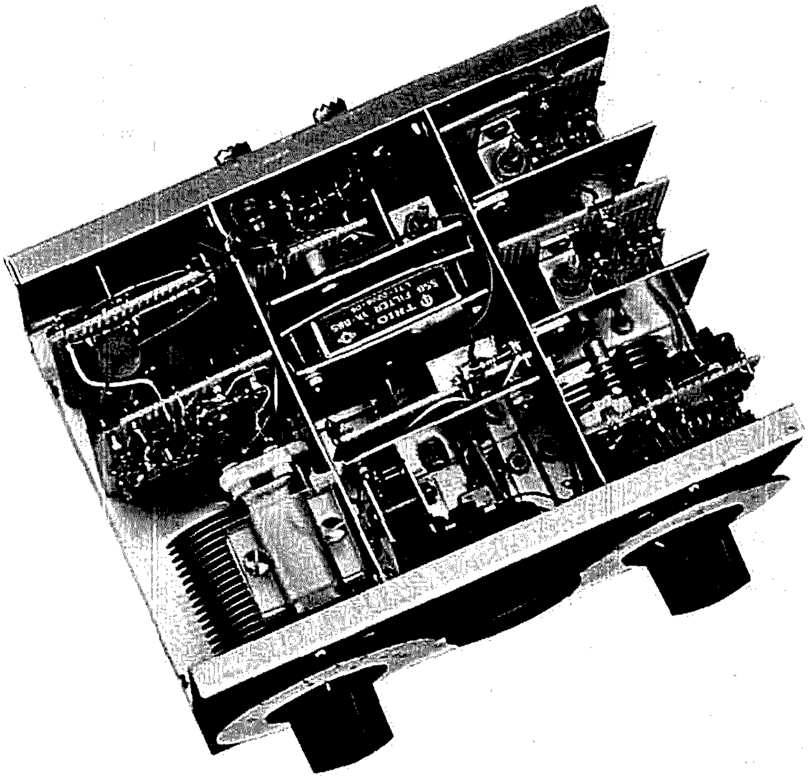
FEBRUARY 1984 / $2.50

ham radio • February 1984

ham radio magazine

- HF receiver performance
- better-sounding SSB
- meteor-scatter communications
- designing elliptic lowpass filters
- EMI/RFI shielding: new techniques (part 2)

THE QUIET FREQUENCY SYNTHESIZER:
USING VXO HARMONIC SELECTION



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contents

cover: W3MT

REFLECTIONS

I enjoy operating because it's relaxing (sometimes), informative ("Roger, you're 5 and 9 — NEXT!"), and a good way to meet interesting people (that's true). It gives me an opportunity to try out new antenna designs, test my understanding of propagation, and, in general, stay more or less informed about what's happening. However, in the process I've noticed a trend worth examining — particularly by those who are part of it. I call it:

INSTANT PUDDING

Recently I've found myself reduced to mumbling into my beard about such-and-such practice by this-or-that individual. It appears that technically, at least, Amateur Radio has come of age, with instant power-on, QSY, tune-up, and antenna matching. We have the latest fully-synthesized, all-mode transceivers, full break-in amplifiers, and motor driven crank-up towers. With even half an effort we can work DXCC in a couple of weeks (or even on a weekend, during a contest). But in the process, something's been left behind. Call it courtesy, experience, operating skill, or just plain consideration for our fellow hams. Let me give you a few examples of what I'm talking about.

WD22XYZ calls CQ, is answered by a DX station, completes the contact, and stands by for any other DX calls on his frequency. Without a moment's hesitation, KD11PQC jumps in with kilowatts a-glowing and calls the DX station. When I was a novitiate to the hobby I learned fast that this is simply not done; unless WD22XYZ relinquishes the frequency, it's still his.

Another pet peeve of mine, heard more times than I care to recall, occurs when a DX station calls a directional CQ: "CQ, CQ, CQ, listening for W1's only!" The ensuing cacophony of calls includes all areas, states, and cities (with only slight exaggeration on my part). Of course, the DX station can control the situation if he continues to request the particular call area and refuses to give in by answering that loud W12 — *maybe*.

Signal quality reporting is another troublesome area. God help the operator who informs another that his signal leaves a little bit to be desired; for example, 20 kHz wide on glorious SSB. I made the mistake of telling another ham the other day that he was transmitting on two different frequencies simultaneously. His immediate reaction was, "That's impossible! It's probably your receiver that's overloading." I hastened to add that *both* of his signals were quite "clean"; it's just that he had two VFO's going at the same time. In all fairness to him, it turned out to have been the other operator in the QSO who was at fault. Fellas — I think most criticism is meant to be constructive. (Anyhow, whatever happened to pride in one's signal/station?)

How many times have you heard, "Roger, roger, roger, you're 5-9 plus. But will you please repeat your call and name and my report?" Please note that Q5 means "perfect copy" and S9, by IARU convention, is 50 microvolts RMS (in a 50 ohm system). By the way, it might be very useful to everyone if you knew what your Brand X receiver S9 was equivalent to. It might make the signal reports a little more meaningful when testing antennas or evaluating the performance of one's station over another's.

Full-power tune-up into the antenna seems to be occurring more and more often these days, usually when you're trying to pull that weak one out. If you *must* tune up on frequency (I won't mention dummy loads), listen first, wait, listen again to make sure that it's a clear frequency and *then* give a short request to see whether the frequency is in use. If it isn't, go ahead. Propagation being what it is, it's quite possible to believe that a frequency's clear when it's really quite occupied. If you have to apply full power to your antenna, load up when the band is truly dead: high noon on 160 meters, 4 A.M. on 10 meters. (I know — 160 and 10 meter operators will insist the bands are *never* dead.)

Communicasting is the term I use to identify operation by an individual who wants others to think he's communicating, even though he's really only broadcasting his highly opinionated, usually insulting, narrow-minded point of view. This practice has reared its ugly head more and more often on the lower bands recently — all in the name of exercising the right to free speech. With all due respect to rights, isn't use of the frequency spectrum — a precious, limited resource — a *privilege*? (If you have something to say, and you want other Amateurs to hear it, write a letter to *ham radio*. If space permits, we'll print it.)

To paraphrase a recent television ad describing the reasons for a stock brokerage's success, courteous operation is just "good old-fashioned work." In our case, if we were to stop expecting instant pudding — i.e., immediate gratification — and instead were more willing to hang in there, learn by example, develop expertise and the kind of operating skills that can only be acquired over a period of time, I, for one, would be a lot happier.

Rich Rosen, K2RR
editor-in-chief

presstop de W9JUV

THE "NO-CODE" LICENSE PROPOSAL WAS SOUNDLY DEFEATED, with not one commissioner voting for it at the December 14 FCC meeting. Presenting the item was Private Radio Bureau Chief Bob Foosaner, who pointed out that while a principal argument for No-Code was the supposed lack of growth in Amateur Radio, the service had grown from 300,000 licensees in 1975 to over 400,000. Furthermore, he continued, the nearly 5,000 comments filed ran 20:1 against No-Code, with even the handicapped it was supposed to benefit opposing it.

Commissioner Mimi Dawson, Studying For An Amateur License Herself, then asked what alternative entry to the radio spectrum might be available for "computer hobbyist" radio communications since Amateur opposition to No-Code did not make Amateur Radio attractive to that group. Foosaner responded, "The hobby class (license) may be something that replaces the CB type of thing...we'll find a spectrum available for this type of service."

Commissioner Comments During The Discussion Were All Complimentary to Amateur Radio, with Chairman Fowler in particular citing the service's many contributions. However, some observers feel the failure of No-Code and the resulting lack of growth in Amateur VHF or UHF band usage will greatly increase outside pressure on those bands. Not only is there the possibility of a non-Amateur "hobby class" competitor for spectrum, but various user groups are actively promoting 216-225 MHz as a land mobile band, plus use of 420-430 MHz.

W5LFL'S STS-9 2-METER OPERATION WAS AN UNQUALIFIED SUCCESS, from every point of view. He operated enough orbits to give the U.S. and the rest of the world at least several chances to hear and—hopefully—"work" him. His "CQ North America" was easily copyable. a thousand miles from Columbia's groundtrack, depending on spacecraft orientation.

Media Coverage Of The Event Was Outstanding, emphasizing the idea that now "the guy next door" could actually talk to an astronaut in space. Owen's chat with King Hussein, JY1, was highlighted, of course, but many average Amateurs across the country found themselves hosting TV news crews and chatting with talk show hosts or newspaper reporters.

Though Owen's Operating Method Disappointed Many who'd hoped to hear their calls come back from space, it made real PR sense. His acknowledging a few calls, then describing what he saw and what was going on in Columbia was far more interesting to non-Amateur listeners and media than long lists of calls. He also reported it almost impossible to copy many calls because of his single-earpiece headset. Though it appears most of those Owen was acknowledging were well-equipped VHF buffs with big signals, Owen's first review of his logging tape after Columbia's return did turn up some mobiles! Discipline was generally fair to good in most areas, though with some confusion and harassment.

Strong, Positive Effects On Amateur Radio are already being felt. Reinforcing the excellent exposure Amateur Radio got during the Grenada invasion, interest in becoming an Amateur seems to be at an all-time high. Inquiries about Amateur training courses are way up, and school-age Amateurs report many classmates asking about their hobby. NASA was also very pleased, with future operation by other licensed astronauts almost certain!

About 300 Callsigns Were Pulled Off The Logging Tape in Owen's initial runthrough. He feels quite a few more can come from a more painstaking review. Of the 300, about 75% were U.S., with no Japanese, Russians or Africans noted. Special QSLs for those who made it on the tape will be available shortly. Send QSLs and SWL reports to ARRL, with an SASE.

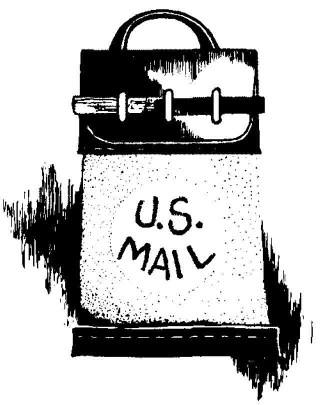
Congratulations to NASA, The ARRL, W5LFL, K6DUE, and many other contributors toward making Owen's space operation possible. A special salute to Vic Clark, W4KFC, who passed away one week too soon to see a dream on which he'd worked so hard finally realized. A most ambitious project, meticulously executed, for the benefit of all—but particularly Amateur Radio.

PCB-FILLED DUMMY LOADS DO POSE A THREAT TO AMATEURS, though perhaps not as severe as once feared. A just-completed study by the Centers for Disease Control in Atlanta found less than 2% of the loads tested contained PCB coolant, though a few others did show traces of the dangerous compound. However, tests of the Amateurs who'd been exposed to the PCB-cooled loads showed no abnormal levels in their blood.

Since Any Exposure To PCBs Is Considered Dangerous, Amateurs who have dummy loads, capacitors, or transformers made before PCB's dangers became known should contact the nearest office of the EPA for advice on safe disposal.

THE VOLUNTEER EXAM IS MOVING AHEAD, with appointment of the first regional Volunteer Exam Coordinator likely very soon. Leading the race with a very well thought out proposal is the Anchorage Radio Club, which is anxious to take on the responsibility for the entire state of Alaska. There are also groups or organizations in almost every other part of the country who've indicated strong interest in taking on the VEC job in their areas. In the meantime, the ARRL is still talking about becoming the national VEC.

WARC FREQUENCY ALLOCATIONS WERE ADOPTED BY THE FCC in December, with several items of concern for Amateurs. 1.9-2.0 MHz was allocated to Amateur Radio on a "restricted secondary" basis, with radiolocation primary. AM broadcast will eventually be moving into 1605-1705 kHz, with displaced radiolocation users then going to 160's top end. 220-225 MHz retains Amateur, plus fixed and mobile pending the results of an FCC NTIA working group study. The 902-928 MHz band seems to have a variety of potential users, but the FCC did turn down industry pleas that it also include a secondary land-mobile allocation.



comments

photovoltaics

Dear HR:

WD8AHO's article, "Photovoltaic Cells: a Progress Report" (December, 1983, page 52), brought to mind painful recollections of the early 1960s when I bought 300 0.4 × 0.8 silicon cells surplus for fifty bucks. I say "painful" because I soldered the little buggers in a series-parallel arrangement. I thought I would go blind and also learned that the little pieces of silicon would fracture easily. When most of the 300 were finally soldered, I had the pleasure of getting all of 2 + watts output on a sunny day. Still, it was great fun after the drudge work of putting it together was done and it did make a nice battery charger. After several years of operation, the unit busted up on my move from San Francisco to Oakland. Too bad — there are more sunshine/hours in Oakland.

The 'breeder concept" photo in the article (**fig. 1**) made me whistle. There must be at least 3000 square feet of panels. That would mean about a *million* of my 0.4 × 0.8 chips (ouch), and at 1960's prices, a cost (even at surplus) of over $150,000, all for about 6 kW output!

Since I've been out of touch with that scene for over twelve years, it's apparent there has been a great deal of progress.

Your article really rekindled my interest.

Nubar Tashjian, K6KVX
Oakland, California

Dear HR:

I don't understand WD8AHO's claim that solar cells at $4.00 per watt will pay back in less than one year. (See "Photovoltaic Cells: A Progress Report," December, 1983, page 52.)

Energy from our local electric company costs $\frac{1W}{1000} \times \$0.052 = \$0.52 \times 10^{-4}$ each hour at the 1 watt rate. So, for the $4.00 cost of a 1-watt solar cell, I could buy 76,923 hours of electric energy at the 1-watt rate. Averaging six hours per day, as one might use a solar cell, I could buy that power from the local electric company for 12,820.5 days or about *35.1 years.*

The added expense for storage batteries (and inverters?) makes the break-even time even longer. I don't know the life expectancy of solar cells, but I would be surprised to break even within the span of human lifetime.

Martin Sample
Tuolumme, California

As Mr. Sample suggests, the question of when photovoltaics will become cost competitive with local power companies is, to some, a good reason not to get involved in photovoltaics.

Photovoltaics is not *for everyone, and certainly not for anyone who measures payback solely in economic terms. But a modest photovoltaics system can meet the needs of the Amateur who wants an economical, uninterrupted supply of power for his or her ham shack, for remote operations, and in emergencies or natural disasters.*

Every day we read of billion-dollar cost overruns, unexpected delays, and unacceptable workmanship in the development of new power plants. These problems suggest that in the years to come, electricity is likely to cost more, not less. It's reasonable to expect that by the time photovoltaic power reaches the 100 megawatt level and the cost has dropped to below $4 per watt, the payback period of PV's — partly because of power company rate increases — will be reduced to less than one year and be competitive with utility rates.

With a life expectancy of 30 years or more, photovoltaics today offer a readily available source of dependable power at a reasonable price.

Paul J. DeNapoli, WD8AHO
Livonia, Michigan

As far as Amateurs today are concerned, perhaps the following letter says it best . . .

Dear HR:

I fail to see how solar energy can have a payback of less than three and one-half years. At $10 per peak watt, and four hours per day peak sun time, I calculate only 1460 watt hours per year. At 7 cents per kWh, this equates to a payback of 10 cents per year — which would take 100 years!

WD8AHO is clearly talking about energy payback only (the energy required to build a cell), not the cost of manufacture. I am truly excited about the prospects of solar energy and finally put cells on my sailboat to keep the batteries trickle-charged. You can't beat PVs when you can't get power any other way, but let's be honest about the cost to the consumer.

R.F. Bruninga, WB4APT
Bowie, Maryland

auto-dialer chip

The MD-22 chip necessary for construction of K2MWU's "State-of-the-Art Auto-Dialer" (December, 1983, page 21) will again be available in mid-to late February.

Readers who have had difficulty in obtaining this part should contact CES, Inc., P.O. Box 2930, Winter Park, Florida 32790. (305) 645-0474.

short circuit

RTTY and the Atari™

In the article, "RTTY and the Atari™ Computer," on page 38 in the July, 1983, issue of *ham radio*, the lower left pinout of the XR2211 chip should be labeled 12. It was omitted from **fig. 3** though discussed in the text on the following page under the subhead "terminal unit construction adjustments."

frequency synthesis
by VXO harmonic selection

Quieter than a PLL, it generates all frequencies between 5 and 6 MHz

This article describes a modification of an old method of frequency synthesis.[1] Its chief advantage over the earlier scheme is that it can synthesize all possible frequencies over the output range, rather than being limited to a set of discrete frequencies. Its stability, determined by the stability of a single VXO, is comparable to that of other synthesizers.

block diagram

The basis for the system is a variable crystal oscillator — a VXO consisting of a 100-kHz frequency standard crystal which is continuously "pulled" from 100 to 100.2 kHz (see fig. 1). The VXO output frequency is divided by ten and fed to a pulse generator. Here a short pulse is formed with a PRF continuously variable over the range 10 to 10.02 kHz, directly controlled by the VXO. A bandpass filter passes the 500th through 600th harmonics of the pulse, producing a "comb" with about 10 kHz spacing which moves continuously upward about 10 kHz as the VXO frequency is increased. If the 500th harmonic is selected, its frequency will move continuously from 5000 to 5010 kHz. Then if the 501st harmonic is selected, its frequency will move continously from 5010 to 5020.02 kHz and so on. In this way we may obtain all possible frequencies in the range from 5 to 6 MHz.

In order to select one harmonic and reject all others, it is necessary to employ a filter having a bandwidth considerably less than 10 kHz. Consideration of spurious frequency problems leads to the use of a SSB crystal filter in the 9 MHz region. I selected an 8.83 MHz filter because I intend to use a 9 MHz beating oscillator to heterodyne the synthesizer into the ham bands, and I do not wish to invite trouble. The VFO and first mixer form a superhet with an 8.83 MHz IF sharp enough to pass just one "tooth" of the "comb." The IF beating against the same VFO in the second mixer restores the selected "tooth" to its original frequency, *exactly*. Drift in the VFO has no effect on the output frequency. As the VFO is tuned continuously in one direction, succeeding "teeth" are selected, interspersed with regions of zero output. The VFO is calibrated in terms of output frequency so that the appropriate "tooth" is selected. The output frequency is then adjusted to the desired value by adjusting the VXO frequency. If this adjustment causes the IF to reject the "tooth," slight retuning of the VFO will bring it back into the filter passband. The selection indicator is a meter which indicates whether or not a "tooth" is passing through the filter. It reads either zero or about 2/3 scale, corresponding to the absence or presence of an output signal, respectively.

theory of operation

Assume a VXO is continuously "pulled" from f_0 to a higher frequency, f_1. We may multiply the frequency by any integer, M, and divide the frequency by any integer, N. The output frequency will vary from

$$f_L = \frac{M}{N} f_0 \tag{1}$$

to

$$f_H = \frac{M}{N} f_1 \tag{2}$$

By Frank Noble, W3MT, 10004 Belhaven Road, Bethesda, Maryland 20817

For continuous coverage, the pulling range of **fig. 2(A)** must be equal to the separation between adjacent harmonics of $\frac{f_0}{N}$, **fig. 2(B)**, or

$$\frac{M}{N}(f_1 - f_0) = \frac{f_0}{N} \quad (3)$$

so that
$$M = \frac{f_0}{f_1 - f_0} \quad (4)$$

Substituting **eq.4** in **eq.1**:

$$N = M\frac{f_0}{f_L} = \frac{f_0^2}{f_L(f_1 - f_0)} \quad (5)$$

Comparing **fig. 2(C)** with **fig. 2(A)**, note that

$\frac{M+1}{N}(f_1 - f_0)$ is larger than

$$\frac{M}{N}(f_1 - f_0) \quad (6)$$

so M in **eq. 4** is the *minimum* value which will provide continuous coverage. Larger values will produce progressively increasing overlap.

practical considerations

If a standard SSB crystal filter is used, the harmonic spacing must be of the order of 10 kHz to allow adequate rejection of neighboring harmonics. Hence

$$\frac{f_0}{N} = 10^4 \quad (7)$$

and
$$N = f_0 \times 10^{-4} \quad (8)$$

Substituting **eq. 8** in **eq. 5**:

$$M = f_L \times 10^{-4} \quad (9)$$

Now the desired value of f_L is 5×10^6, so that

$$M = 500 \quad (10)$$

AT-Cut crystals will not achieve values of M as small as 500, using capacitance tuning. This amount of pull may be obtained with inductive tuning, but extreme care must be taken, especially with the coil, to retain stability. The frequency, f_0, must exceed 10 MHz so that reasonable values of inductance may be used. This leads to the requirement of large frequency division, N, an additional complication. An incidental advantage is that the tuning linearity is better than for capacitance loading.

If we use the 100 kHz 5 degree X Cut commonly employed for frequency standards, a value of $M = 500$ may be obtained with capacitance tuning. The temperature stability will be poorer than for AT crystals, but a crystal oven will solve this problem.

In any case, the percentage drift of the output frequency will be identical to the percentage drift of the crystal.

circuit description

Fig. 3 shows a detailed circuit diagram for one form of the synthesizer.

The first stage, the VXO, uses an LF353N for U1A in a power design[2] and substitutes a large variable capacitor having "midline" plates to improve the tuning linearity. (SLF or even more radical plate shapes would be preferable if available.) A fast-rising square wave is formed by U1B, suitable for driving the divide by ten IC, U2. The very fast rising square wave output of U2 is differentiated by a short time-constant RC circuit. The negative-going pulse is clipped by the

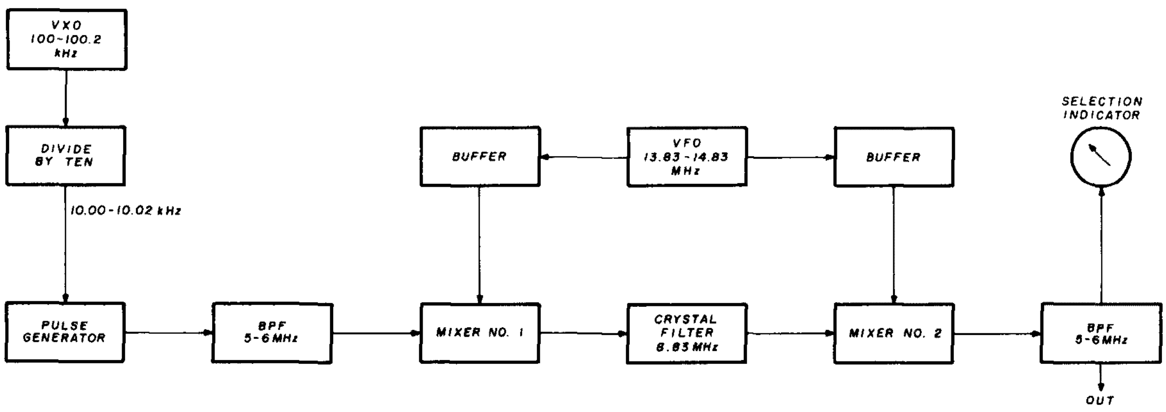


fig. 1. One possible arrangement for the synthesizer. See text for a discussion of the relationships between the various frequencies.

diode, so that the input to Q1 is a short positive pulse with a PRF continuously variable from 10 to 10.02 kHz. The source of Q1 is biased at +6 volts so that only the narrow part of the pulse near its peak is amplified. The tuned circuit in the output of Q1 is resistively loaded to pass 5-6 MHz, selecting the desired "comb." Q2 raises the comb level. Its output is also broadbanded by resistive loading. A link on the Q2 output coil drives one input of diode ring mixer No. 1 through an attenuator which provides a 50-ohm source for all frequencies. The second input to the mixer is supplied by the VFO buffer, Q9, having a source impedance of about 50 ohms at the VFO frequency. (This is not ideal. It would have been better to increase the output of Q9 and use an attenuator to provide 50 ohms at all frequencies.)

The output of the first mixer is terminated in 50 ohms for all frequencies. *This is the most critical termination.* It drives FL-1 through a series resistor, forcing a termination of less than 600 ohms, which is roughly what the filter requires. The output termination is about 400 ohms. (We observe that passband flatness, vital to SSB, is not important in this application; consequently, termination requirements have been eased.) Since the filter is very sharp, the output circuit of IF amplifier U3 has high Q and large L/C to achieve high gain. Q3 is an RC amplifier for driving FL-2 with a source impedance of somewhat less than 680 ohms. This filter is also terminated in somewhat less than 680 ohms. It drives emitter-follower Q4.

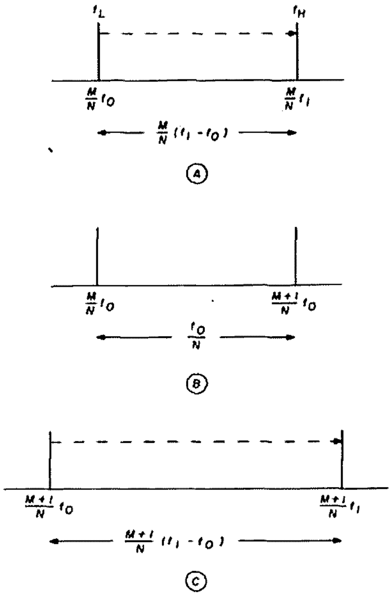
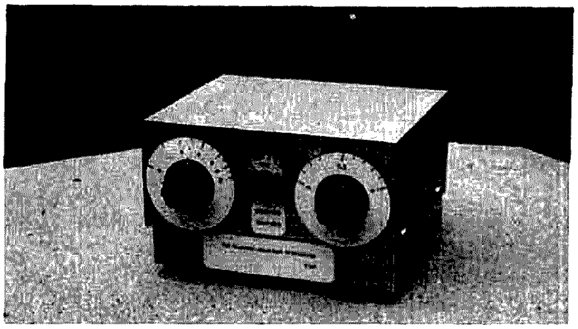


fig. 2. Diagram for determining the relationships between frequencies, multipliers, and divisors.



Left, VXO direct drive dial calibrated in kHz. Center, selection indicator meter. Right, VFO reduction drive dial calibrated in 100 kHz steps.

The output impedance of Q4 approximates 50 ohms at all frequencies to match one input of the second mixer. A separate VFO buffer, Q8, drives the other input. The critical output of the second mixer sees 50 ohms at all frequencies. U4 is an amplifier, broadbanded by the 50 ohm half-wave filter at its output. The input to amplifier U5 is fairly clean. The signal is made larger by U5 and further laundered by the bandpass filter at its output. Q5 is an emitter follower to reduce the output impedance to about 50 ohms at the "HI" output. An L network followed by a quarter-wave filter reduces the output voltage and impedance and further improves the waveform at the "LO" output. Q6 is another emitter follower used to prevent distortion generated by the diode rectifier from reaching the RF outputs. The rectified and filtered output of Q6 is used to drive the selection indicator, a miniature 1 milliampere meter. The VFO, Q7, is a Hartley oscillator using a good coil and source degeneration to obtain good stability and waveform. It must not drift over ±1 kHz per hour — otherwise it could drift the desired "tooth" out of the filter slot. The VFO tuning capacitor has semi-circular plates for good tuning linearity. It is driven by a reduction dial to make the tuning less critical.

mechanical details

Except for the portion between the U4 link and the outputs, the circuit is contained in a metal enclosure (standard chassis No. 1) measuring 3 1/4" H × 7 1/2" W × 5 7/8" D, (9 cm × 19 cm × 15 cm) Radio Shack No. 270-229 (see photo 2). A second standard chassis (No. 2) contains the portion after the U4 link, a 12-volt regulated power supply (150 mA), and a converter employing a 12.5 MHz crystal oscillator to beat the synthesizer output up into the 40-meter band. No attempt was made to miniaturize chassis No. 2 because it rests on a shelf under the operating table, out of the way. The coils in chassis No. 2 are surplus 5/8-inch slug-tuned bakelite units mounted in shield

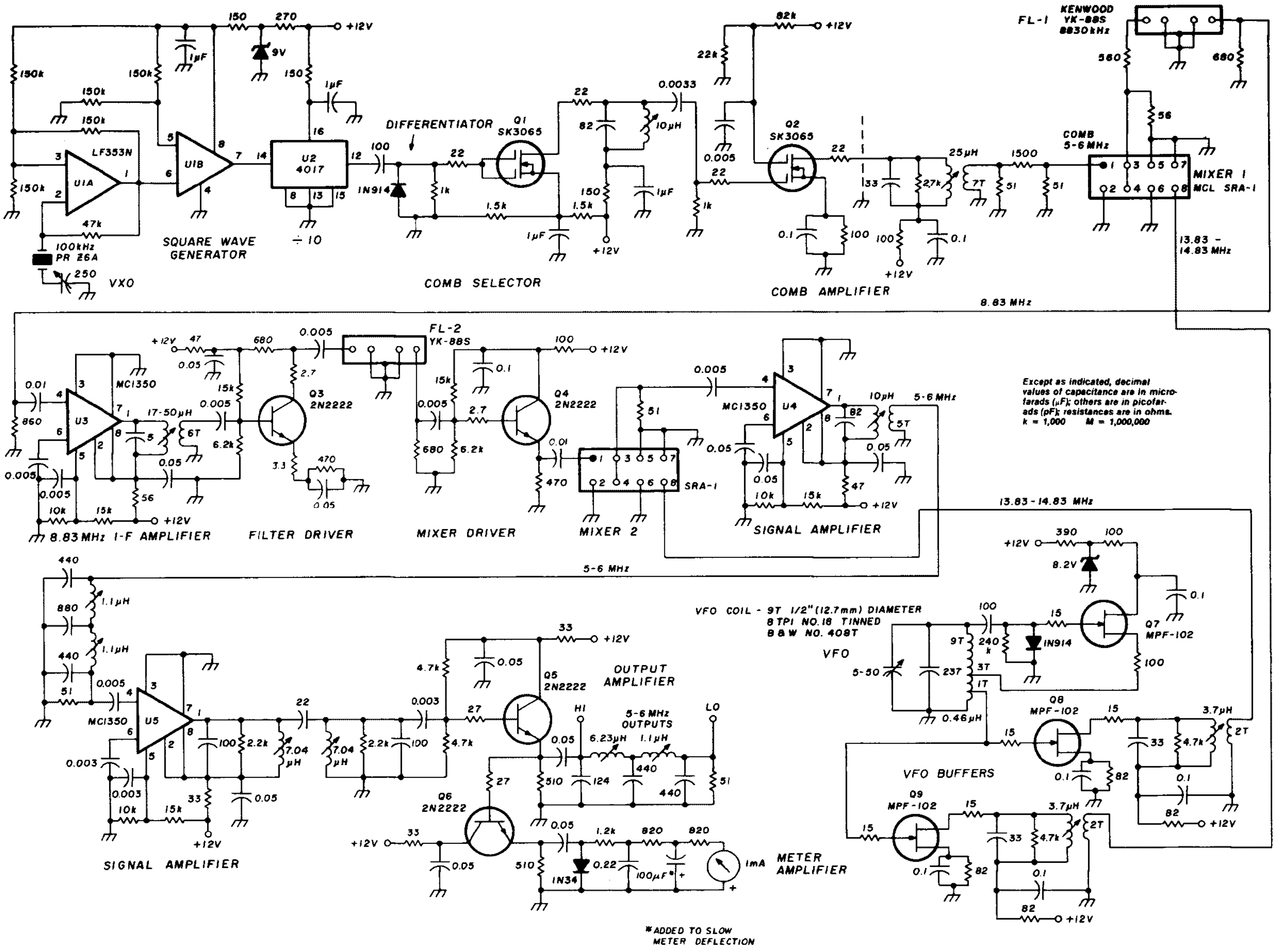
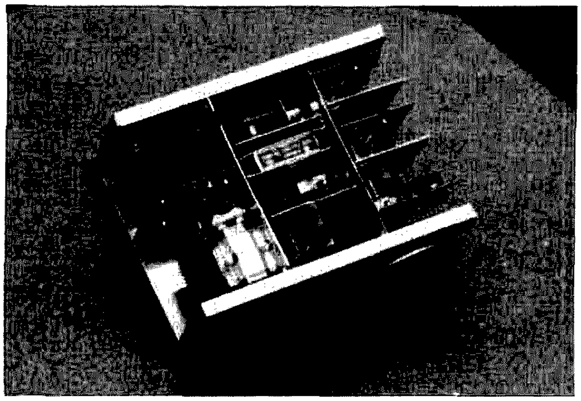


fig. 3. Detailed circuit diagram for one form of the synthesizer. Details of the design and suggestions for improvement are contained in the text.



Left, front-to-back, VXO through comb amplifier. Right, front-to-back, VFO and VFO buffers. Center, balance of circuitry through U4 output; front-to-back, Mixer No. 1, FL-1, U3, Q3, FL-2, Q4, Mixer No. 2, U4.

cans. The coils in chassis No. 1 are 3/8-inch slug-tuned ceramic unshielded forms made by Miller. A shielded cable containing power supply leads and the DC for the selection indicator interconnects the two chassis. The RF interconnection is made by means of coax cable.

Chassis No. 1 is divided into three main shielded compartments by means of two aluminum shields running front-to-back. The left compartment contains U1 through Q2. The right compartment is subdivided into three shielded sections and contains Q7, Q8, and Q9, front-to-back. The remainder of the circuit through U4 is contained in the center section. Most of the circuits are supported on perfboard placed parallel to the front panel. Push-in terminals are used where required. Care is taken to be sure that the mixer grounds are of low inductance and resistance — otherwise balance will be degraded. Similarly, the connections to the crystal filter grounds and case should be very solid, and the input and output circuits well separated and shielded; otherwise, unwanted signals may bypass the filters. The VXO tuning shaft is directly driven by a knob calibrated in kHz. The VFO shaft is driven by a 6:1 reduction drive dial calibrated in 100 kHz steps. Coax cable running below the bottom plate carries the RF interconnections between the shielded compartments. For mechanical reasons, the ends of the coax are exposed. To correct this, a shield resembling an inverted chassis is placed below the bottom plate and secured with metal spacers. This shield also raises the knobs above the table, increasing the ease of tuning.

comments

The writer's experience with low frequency VXO circuits is limited to one crystal of ancient manufacture. Before proceeding further, a prospective builder should experiment to be sure his 100 kHz crystal will pull 200 Hz. (Note that the exact fre-

quencies do not matter; the difference between the maximum and minimum frequencies must be at least 200 Hz.) If this much pull cannot be obtained, the circuit will have to be re-designed using the equations contained above.

It should be clear that this is a first model which is certainly not of optimum design. Its defense is that it presents no surprises. The second crystal filter, used to reduce noise generated in the IF amplifier and to increase the out-of-band attenuation, may not be needed. The filters associated with U5 and Q5 could be simplified and improved. The crystal could be placed in an oven for better stability. Finally, the VFO reduction drive should have a higher ratio, (preferably 100:1, to ease the tuning); if not, some form of electrical bandspreading could be used.

acknowledgements

I appreciate the interest in this project expressed by many fellow hams. In particular, I wish to express my gratitude to Dick Schellenbach, W1JF, and Skip Marsh, W6TFQ, for their help and encouragement.

references

1. E.W. Pappenfus, "Electronics for Communication Engineers," *Markus and Zeluff*, McGraw-Hill, first edition, 1952, pages 377-378.
2. "Frequency Measurements," *ARRL Handbook*, American Radio Relay League, 1982, page 16-13.

ham radio

elliptic lowpass audio filter design

Use standard 88 and 134 mH inductors for deep attenuation filter response

The elliptic lowpass filter is frequently used by the Radio Amateur for speech filtering because of its sharp selectivity, and many articles have been published on this particular application.[1-6] Unfortunately, the audio filter design procedures currently used by the Amateur are derived from engineering texts, and therefore are more suited to commercial practices. For example, the commercial filter designer develops his design based on a particular cutoff frequency and impedance level, and then orders the inductors from an inductor manufacturer. Capacitors are obtained from electronic distributors, measured, and then paralleled as required to get the design values within the specified tolerance. In Amateur applications, in which requirements are less stringent, variations in cutoff frequency and impedance level can be exchanged for more convenient component values. The Amateur also places greater importance on being able to use an inexpensive source of standard-value inductors than does the commercial designer. This is because the Amateur usually builds only one or two filters and therefore cannot take advantage of the large-volume discount prices available to the commercial designer.

In the discussion that follows, a simple design procedure for passive LC audio elliptic lowpass filters that use inexpensive 88 and 134-mH inductors is presented.

common filter types

The two most frequently used filter types in Amateur Radio practice are the Chebyshev and the elliptic. Both can be designed for any level of passband ripple; the elliptic is more versatile because it also can be designed for any level of stopband attenuation.[7] Also, the attenuation rise of the elliptic can be made much more abrupt than the Chebyshev.

In RF harmonic filtering, attenuation greater than 40 dB one octave from the cutoff frequency is usually adequate; a 7th-degree Chebyshev design is most suitable because its response provides more than 42 dB attentuation at twice the cutoff frequency and it has a simple ladder configuration (four shunt capacitors alternating with three series inductors). However, for speech filtering applications, it is important that the filter attenuation rise as quickly as possible to 40 dB or more. Consequently, the 7th-degree elliptic design is preferred over the Chebyshev. Greater than 40 dB of attenuation occurs at a frequency only 1.2 times the cutoff frequency of an elliptic filter compared to two times the cutoff frequency of a Chebyshev design.

Fig. 1 shows the schematic diagram and the typical attenuation response of a 7th-degree elliptic filter. A family of elliptic designs is available that shows 45 to 65 dB of stopband attenuation and uses the *same value of inductance for L4 and L6.* This means that only two different inductance values are needed (L2, and L4, L6). One value can be the fixed 88 or 22-mH value of the standard surplus inductor, and the other can be obtained from a modified 134 or 33.5-mH inductor.

inductor types

The two inductor types suitable for the elliptic filter construction are shown in **fig. 2.** Both types (one potted) have molybdenum-permalloy cores and the optimum frequency range is from 300 Hz to 20 kHz. Both types have two windings that can be connected either in parallel or series aiding. **Fig. 3** shows the inductors with instructions for connecting the windings in series or parallel.

The unpotted inductor has an inductance of 134 or 33.5 mH in the series or parallel connection, respectively. The potted inductor has values of 88 or 22 mH in the series or parallel connection. The potted induc-

By Ed Wetherhold, W3NQN, Honeywell, P.O. Box 391, Annapolis, Maryland 21404

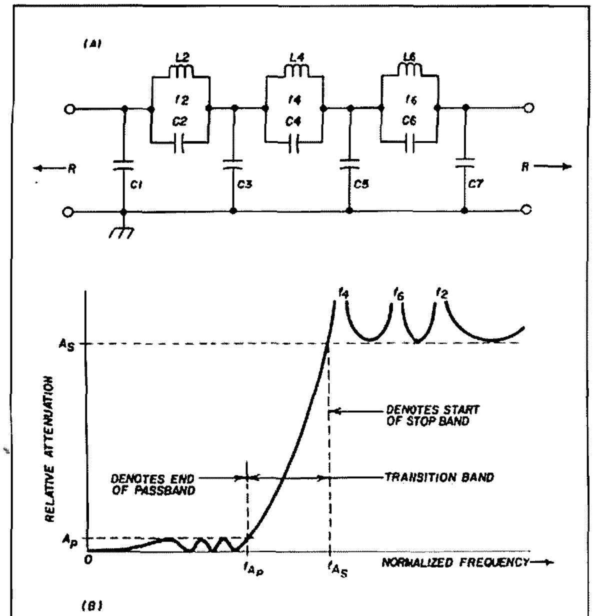


fig. 1. Elliptic 7th-degree lowpass filter. (*A*) Schematic diagram including component designations. (*B*) Attenuation response.

tor is designed for mounting on a flat surface by inserting a ¼-inch diameter stud through a hole and then by placing a Tinnerman mounting clip over the stud to secure the inductor. The center inductor in **fig. 2** is shown with a Tinnerman mounting clip.

Since turns cannot be removed from the potted inductor to vary the inductance, it must be used as is in applications requiring a 22 or 88-mH value. Because turns can be removed from the unpotted inductor, many different elliptic designs are possible, and eight different designs have been selected and calculated for stopband attenuation (A_s) levels ranging from 45.6 to 64.5 dB. The total turns needed to be removed from the unpotted inductor for these eight designs varies from eight to one hundred turns. Procedures and equations for calculating the total turns to remove for any inductance value that may be needed are given in **appendix A.**

7th-degree elliptic lowpass filter

The maximum amplitude of the passband ripple (A_p) (see **fig. 1B**) is less than 0.11 dB for these designs. The cutoff frequency is designated f_{A_p} and is defined as that frequency at which the attenuation first exceeds the A_p level. The normalized value of f_{A_p} is unity, and all other frequencies are referenced to it. The start of the stopband is denoted by f_{A_s} and occurs where the minimum stopband attenuation (A_s) is first achieved. The band of frequencies between the end of the passband and the start of the stopband is called the transition band. It is desirable to minimize the transition band for best filter selectivity. An indication of the degree of selectivity is the f_{A_s}/f_{A_p} ratio — the smaller the ratio, the more selective the filter. Since the normalized value of f_{A_p} is unity, the normalized value of f_{A_s} indicates filter selectivity. The f_{A_s} values in the eight designs vary from 1.2 to 1.37. Note: selectivity and higher stopband attenuation are tradeoffs.

In audio filtering applications, a passband ripple of less than 0.2 dB (as found in the eight designs), is sufficiently low and won't be discussed further.

L2-C2, L4-C4, and L6-C6 resonant circuits produce attenuation peaks in the response at f_2, f_4, and f_6 (see **fig. 1**). If these attenuation peaks are correctly placed, the minimum stopband attenuation will theoretically never drop below the A_s value.

simplified design procedure

Table 1 lists all filter parameters: inductances, reflection coefficients, A_p, f_{A_s}, and A_s. The reflection coefficient, used by filter designers to categorize a specific design, is directly related to A_p. The relationship between R (filter termination impedance) and f_{A_p} is given in columns R and F of **table 1**. R can be calculated if f is known, or vice versa. The L2 and L4, L6 columns list two inductance values which correspond to the series or parallel connections of the inductor windings.

Table 2 lists the normalized component values and frequencies of the same eight designs in **table 1**. This data is used to calculate the actual component values and frequencies of the filter after the termination resistance and cutoff frequency are selected.

The designs in **tables 1** and **2** are keyed to each other either by the design number or by the *L2/L4*

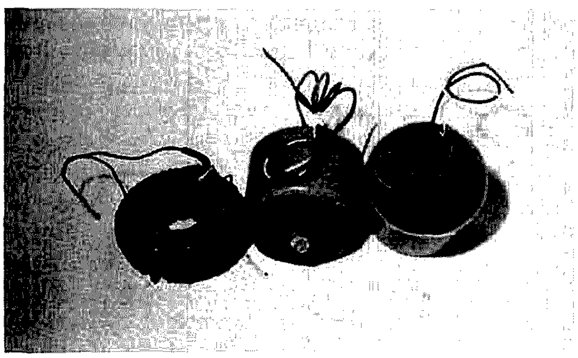


fig. 2. The two inductor types used in constructing the filter. Each has two windings which can be connected in series or parallel. The large unpotted inductor (L2) has a maximum inductance of 134 or 33.5 mH and the potted inductors (L4 and L6) have 88 or 22 mH for the series or parallel connections, respectively. All inductors are shown with their leads in the parallel connection. One of the potted inductors is shown with a Tinnerman mounting clip.

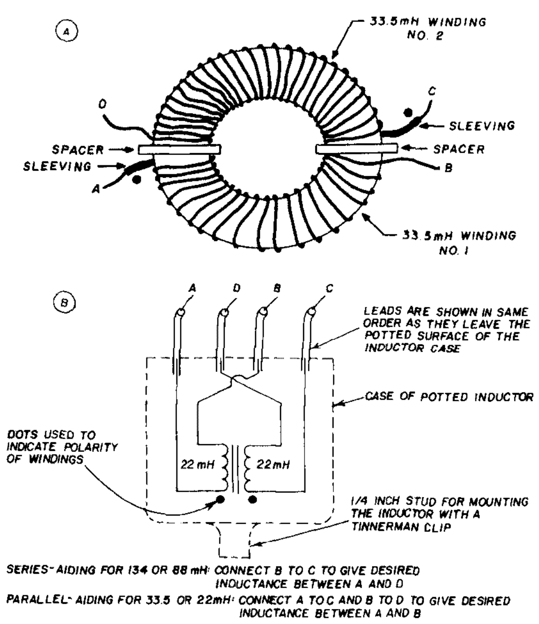


fig. 3. These diagrams show the lead designations of the unpotted 134 mH and potted inductors with instructions for connecting the two windings in series or parallel: (A) unpotted 134 mH; (B) potted 88 mH.

ratio. These ratios range from 1.25 to 1.50, and correspond to $L2$ values from 110 to 132 mH (in the series-aiding connection) or from 27.5 to 33.0 mH (in the parallel-aiding connection). The purpose of each table will be explained before demonstrating the design procedure.

Before selecting a design from **table 1**, the cutoff frequency and the approximate impedance level must first be selected. Because the cutoff frequency is usually more critical, its value is fixed, and R is calculated using the equations under column R or by using the graph in **fig. 4**. The "A" designs are best suited for cutoff frequencies below 2 kHz, while the "B" designs are best for cutoff frequencies above 2 kHz.

In most cases, an exact match between design and circuit impedance is not possible. However, resistive matching pads (see **appendix C**) can be used if the difference between the circuit and filter impedances is less than about 30 percent. For greater impedance differences, matching transformers must be used. The disadvantage of not being able to select a specific and independent filter termination impedance is a consequence of using fixed inductor values, worthwhile because standard-value surplus inductors can be used. The filter performance parameters are next reviewed to see if they are fully satisfactory. If, for example, a minimum stopband attenuation exceeding 55 dB is required, then only designs 1 through 3 should be considered. After a design is selected from **table 1**, and normalized values of the same numbered design in **table 2** are scaled to the impedance level and cutoff frequency.

design example

To design a speech filter that has a 3 kHz cutoff frequency and a source and load impedance of 500 ohms, a typical procedure would consist of the following steps:

(a) Follow the 500-ohm line across the graph in **fig. 4** until it intersects with the slanted lines in the vicinity of 3 kHz; design 8B provides the closest match. If a lower or higher cutoff frequency is required, the filter impedance level must be changed to correspond to the new cutoff frequency. If the new filter impedance is within 30 percent of 500 ohms, resistive padding can be used to match the filter to the 500-ohm system. For greater impedance differences, suitable matching transformers must be used to prevent excessive signal amplitude loss (see note 1).

(b) Refer to **table 1** to check the performance parameters of design 8B. If f_{A_S} and A_S are satisfactory, then design 8B can be used. For design 8B, $L4 = L6 = 22\ mH$, $L2 = 33\ mH$, $L2/L4 = 1.500$, and $R.C. = 8.06$ *percent*.

(c) Find the exact f_{A_p} cutoff frequency for $R = 500$

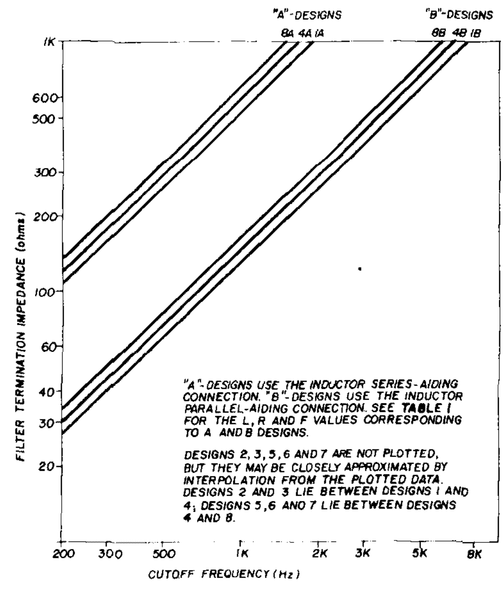


fig. 4. Lowpass filter cutoff frequency versus termination impedance for some of the filter designs in *table 1*.

ohms from the equation in column f of **table 1**: $f = 6.0137(500) = 3007\ Hz$. Use 3007 Hz as the cutoff frequency and $R = 500\ ohms$ in the following calculations. (By a fortunate coincidence, in this particular case both the f and R values are virtually identical with the desired values. This will not always be the case, and it is up to the ingenuity of the filter designer to find an optimum combination of parameters that best satisfies the design requirements. This is one of the things that makes filter designing a challenge.)

(d) Calculate the capacitance and inductance scaling factors, C_S and L_S where R and f are in ohms and Hertz:

$$C_S = 1/(R \cdot f) = 1/(500)(3007) = 0.6651 \cdot 10^{-6}$$
$$L_S = R/f = 500/3007 = 166.3 \cdot 10^{-3}$$

(e) Calculate the L and C component values by multiplying the normalized component values of design 8, **table 2**, by the C_S and L_S scaling factors. Frequencies are obtained by multiplying the normalized frequencies of design 8, **table 2**, by the cutoff frequency, 3007 Hz. The component and frequency values obtained are:

C1	$= 0.1289f\ (0.6651 \cdot 10^{-6}) = 0.0857\ \mu F$
C2	$= 0.02561f\ (0.6651 \cdot 10^{-6}) = 0.01703\ \mu F$
C3	$= 0.2084f\ (0.6651 \cdot 10^{-6}) = 0.1386\ \mu F$
C4	$= 0.1291f\ (0.6651 \cdot 10^{-6}) = 0.0859\ \mu F$
C5	$= 0.1819f\ (0.6651 \cdot 10^{-6}) = 0.1210\ \mu F$
C6	$= 0.09836f\ (0.6651 \cdot 10^{-6}) = 0.0654\ \mu F$
C7	$= 0.08079f\ (0.6651 \cdot 10^{-6}) = 0.0537\ \mu F$
L2	$= 0.19845H\ (166.3 \cdot 10^{-3}) = 33.0\ mH$
L4,6	$= 0.1323H\ (166.3 \cdot 10^{-3}) = 22.0\ mH$
f_{A_p}	$= 1.000\ (3007\ Hz) = 3007\ Hz$

table 1. Lowpass filters: inductor values and L2/L4 ratios with related parameters for selected 7th-degree elliptic designs where $L4 = L6$. [For filter construction using two potted 88-mH inductors and one modified 134-mH inductor.]

design no.		L2/L4 ratio	L2 (mH)	L4,6 (mH)	R.C. (%)	A_p (dB)	f_{A_s}	A_s (dB)	R (ohms)	f (Hz)
1	A	1.250	110.0	88	15.36	0.1037	1.3703	64.50	0.5279(f)	1.894(R)
	B		27.5	22					0.1320(f)	7.577(R)
2	A	1.300	114.4	88	12.80	0.07171	1.3166	59.16	0.5515(f)	1.8133(R)
	B		28.6	22					0.1379(f)	7.2532(R)
3	A	1.333	117.3	88	11.56	0.05842	1.2888	56.19	0.5689(f)	1.7578(R)
	B		29.3	22					0.1422(f)	7.0313(R)
4	A	1.380	121.4	88	10.23	0.04571	1.2573	52.61	0.5947(f)	1.6817(R)
	B		30.4	22					0.1487(f)	6.727 (R)
5	A	1.400	123.2	88	9.77	0.04164	1.2458	51.25	0.6060(f)	1.6501(R)
	B		30.8	22					0.1515(f)	6.6005(R)
6	A	1.440	126.7	88	8.98	0.03517	1.2258	48.77	0.6293(f)	1.5890(R)
	B		31.7	22					0.1573(f)	6.356 (R)
7	A	1.460	128.5	88	8.64	0.03257	1.2169	47.64	0.6411(f)	1.5597(R)
	B		32.1	22					0.1603(f)	6.239 (R)
8	A	1.500	132.0	88	8.06	0.02829	1.2012	45.56	0.6652(f)	1.5034(R)
	B		33.0	22					0.1663(f)	6.0137(R)

Notes:

1. The unpotted inductor (L2) has two 33.5 mH windings that can be connected in series or parallel aiding to give 134 or 33.5 mH, respectively. The potted inductors (L4 and L6) have two 22 mH windings which can be connected in series or parallel aiding to give 88 or 22 mH. **Fig. 3** shows the series and parallel aiding connections.

2. See **fig. 1A** for the schematic diagram and **fig. 1B** for the typical attenuation response of the above filter designs. See **table 2** for the normalized component values and frequencies.

3. To obtain the L2 inductance values, remove an equal number of turns from both windings of the unpotted inductor in accordance with the following equations (as shown in **appendix A**):

(a) for all "A" designs, $T_s = 532 - 45.957\sqrt{L_s}$, where T_s is the number of turns removed from each winding (total turns removed $= 2 \cdot T_s$, and L_p = the desired L2 inductance in the series-aiding connection.

(b) for all "B" designs, $T_p = 532 - 91.914\sqrt{L_p}$, where T_p is the number of turns removed from each winding (total turns removed $= 2 \cdot T_p$), and L_s = the desired L2 inductance in the parallel-aiding connection.

For example, for L_s or L_p = 132 or $33\ mH$, remove four turns from each winding and connect the windings in the series or parallel-aiding connection, respectively.

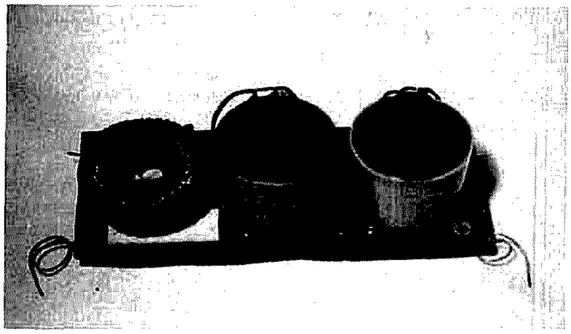
f_{A_s} = 1.201 (3007 Hz) = 3611 Hz
f_2 = 2.233 (3007 Hz) = 6715 Hz
f_4 = 1.218 (3007 Hz) = 3663 Hz
f_6 = 1.395 (3007 Hz) = 4195 Hz
with R = 500 ohms and A_s = 45.6 dB

As a check on the correctness of the values of (C2,L2), (C4, L4) and (C6,L6), the resonant frequencies of these parallel tuned circuits can be calculated and should agree to within about 0.1 percent with the values of f_2, f_4, and f_6 calculated from the scaling procedure.

filter construction

Figs. **5A** and **5B** are photographs of the completed filter design 8B. The inductors and capacitors are mounted on opposite sides of a 2 x 5-inch terminal board. The unpotted inductor is fastened to the terminal board with 3M Scotch™ tape, and the two potted inductors are mounted with their 1/4-inch mounting studs and Tinnerman clips.

The tolerances of the shunt capacitors (C1, 3, 5, and 7) are not critical, and values within 5 percent of the design values are satisfactory. Two capacitors can be paralleled when necessary to obtain the required value. The calculated values of the resonating capacitors (C2, 4, and 6) should be used only as a guide. This is because the actual resonant frequencies of the three tuned circuits are of primary importance, while the exact values of capacitance are of secondary importance. The 88-mH value of the potted inductors is only nominal, and it can vary as low as 86 mH. It seldom is greater than 88 mH. The winding capacities of the potted inductors in the 88- and 22-mH connections are about 68 and 176 pF, respectively. The winding capacities of the unpotted inductor in the 134- and 33.5-mH connections are about 68 and 148 pF, respectively. The simplest procedure for tuning the three resonant circuits of the elliptic filter is to assemble all the filter com-


(A)

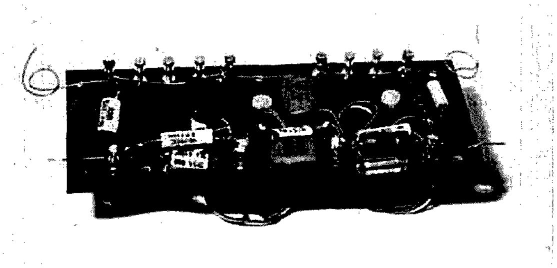

(B)

fig. 5. The photographs of the completed filter show how the inductors and capacitors are mounted on opposite sides of a standard terminal board. The filter assembly can then be mounted in a standard 5 × 4 × 3-inch aluminum mini-box. In (*A*), the unpotted inductor (L2) is on the left followed by L4 and L6. In (*B*), capacitor C1 is on the left. The common ground return wiring is at the top of the terminal board.

table 2. Lowpass filter: normalized component values and frequencies of selected 7th-degree elliptic designs where *L4* = *L6*. (Values are normalized for an f_{A_p} cutoff frequency of 1 Hz and 1-ohm terminations.)

design no.	L2/L4 ratio	C 1 (Farad)	C 2	L 2 (Henry)	C 3 (Farad)	C 4	L4&L6 (Henry)	C 5 (Farad)	C 6	C 7	f_{A_s}	f_2	f_4	f_6
1	1.250	0.1760	0.01627	0.2083	0.2671	0.07819	0.1667	0.2439	0.05647	0.1438	1.370	2.734	1.394	1.640
2	1.300	0.1611	0.01832	0.2074	0.2495	0.08863	0.1596	0.2251	0.06487	0.1254	1.317	2.582	1.338	1.564
3	1.333	0.1535	0.01964	0.2063	0.2402	0.09552	0.1547	0.2152	0.07047	0.1155	1.289	2.501	1.309	1.524
4	1.380	0.1448	0.02141	0.2042	0.2293	0.1051	0.1480	0.2036	0.07830	0.1038	1.257	2.407	1.276	1.478
5	1.400	0.1416	0.02214	0.2033	0.2252	0.1091	0.1452	0.1993	0.08165	0.09931	1.246	2.372	1.264	1.462
6	1.440	0.1360	0.02357	0.2014	0.2179	0.1172	0.1398	0,1916	0.08835	0.09126	1.226	2.310	1.243	1.432
7	1.460	0.1335	0.02426	0.2004	0.2145	0.1211	0.1373	0.1881	0.09169	0.08758	1.217	2.283	1.234	1.419
8	1.500	0.1289	0.02561	0.19845	0.2084	0.1291	0.1323	0.1819	0.09836	0.08079	1.201	2.233	1.218	1.395

Notes:

1. **Fig. 1A** shows the interconnections of the capacitors and inductors. **Fig. 1B** shows the cutoff frequency and attenuation peaks corresponding to the f_1 and $f2$, $f4$, and $f6$ values.
2. The normalized data for the filter design numbers in this table also apply to similarly numbered designs in **table 1**.

ponents except for the resonating capacitors, C2, C4 and C6. Connect the partially assembled filter between a signal source and a resistive load, both having the design impedance of the filter. Set the signal source to the frequency of the first resonant circuit to be tuned, and monitor the filter output voltage across the load resistor with an AC VTVM. Resonate the inductor of the first resonant circuit with a calibrated capacitor decade. Resonance will be indicated by a minimum reading on the AC VTVM. Read the capacitance required for resonance from the capacitor decade, and then place this value of capacity across the inductor using as many paralleled capacitors as required. Repeat the same procedure for the other two resonant circuits. Using this procedure, it is possible to tune the resonant circuits to within 0.25 percent of the design frequency values.

For optimum stopband performance, the ground leads of C1, 3, 5, and 7 must be independent of each other to minimize any common ground impedance.

testing the filter

Fig. 6 shows the measured filter attenuation compared with the computer-calculated response. Close agreement between measured and calculated responses in the pass and transition bands indicates that a good filter design and assembly procedure were used. The lower-than-expected attentuation peaks at f_2, f_4, and f_6 are because of a common ground impedance in the ground leads of the four shunt capacitors. If separate ground leads are used from the shunt capacitors, the peak attenuation levels can be increased by more than 10 dB. This will cause the measured minimum attenuation A_S levels to increase slightly.

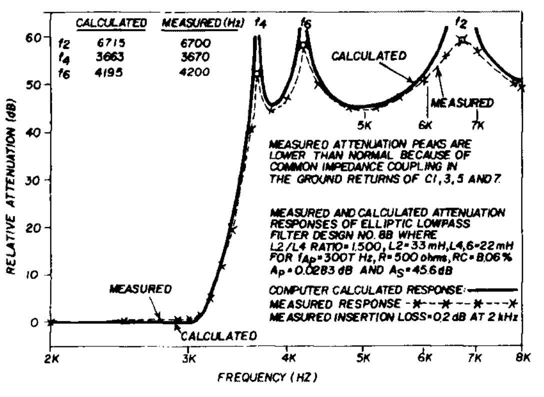


fig. 6. Calculated and measured attenuation responses of the 3-kHz filter design example.

The passband insertion loss is only 0.2 dB. This low value is due to the relatively high-Q inductors in the 3 kHz to 8 kHz frequency range. For much lower cutoff frequencies, the inductor Q's will be lower and the insertion loss will be higher.

obtaining surplus inductors

The potted 88-mH inductors have been provided to the author at no charge by the C & P Telephone Company of Maryland for distribution to those Radio Amateurs wishing to build audio filters. The 134-mH inductors were purchased by the author and are being offered at cost to those wishing to build the filter designs described in this article. Your request for inductors should be addressed to me and accompanied by a description of your filtering application and a 4 × 9-1/4 inch SASE.

summary

Until now, the Radio Amateur has been hindered in using high-performance audio lowpass filters due to the lack of a simplified design procedure and a source of high-Q and inexpensive inductors. This article explains how these problems have been solved. By using computer-aided calculations, designs from a special family of elliptic lowpass filters have been derived that can use two surplus potted 88-mH inductors and one unpotted 134-mH inductor.

Data was tabulated for eight 7th-degree elliptic lowpass designs suitable for the audio frequency range, and a simplified design procedure was demonstrated by designing and building a 3-kHz, 500-ohm lowpass filter. Comparisons were made between the measured and computer-calculated responses to show that the design and construction procedures were valid. It was explained how these design procedures usually result in non-standard impedance levels, and how this inconvenience can be alleviated by using padding resistors or inexpensive matching transformers. A source of surplus inductors for the filter construction was given, and a procedure was explained for deriving an equation for modifying the unpotted inductor for a specific inductance.

acknowledgement

The author gratefully acknowledges the assistance of Mike Barge of Honeywell for performing the computer calculations required in deriving the elliptic filter designs used in this article. The author is also grateful to Joseph Gutowski of EWC, Inc., Harold Mitchell, N0ARQ, of 3M Electronic Products Division, Rex Cox, formerly of Honeywell, and Frank Noble, W3MT, for their review of and comments on the article manuscript before it was submitted for publication.

appendix A

For a series-aiding connection calculate the total number of turns, T_o, as follows:

(a) Measure the original inductance, L_o, with the windings connected in series-aiding (SA) using an inductance bridge. An indirect method of inductance measurement is to resonate the inductor with a precisely known capacitance (± 0.5 percent) while measuring the resonant frequency with a digital frequency counter.[9] Calculate the inductance from the equation:

$$L_{(mH)} = 25.33/(f^2C) \qquad \textbf{(1)}$$

where f and C are in kHz and μF.

Use coupling capacitors of approximately 500 pF each to isolate the tuned circuit from the signal generator and the AC VTVM.

(b) Remove 50 turns from each winding and connect the windings in series aiding. Measure the modified inductance, L_r.

(c) Calculate $T_o = 100 \cdot R/(R - 1)$ **(2)**

where $R = \sqrt{L_o/L_r}$
and T_o = original number of turns on the unmodified inductor
L_o = original inductance (mH) in the series-aiding (SA) connection
L_r = inductance (mH) for the SA connection after 100 turns were removed

For example, if $L_o = 134\ mH$ and $L_r = 110\ mH$ for 50 turns removed from each winding, then $R = 1.10371274$ and $T_o = 1064$.

The unique value of T_o is used to find the equation that gives the total number of turns to be removed from any inductor (of the same type) in order to obtain a specific inductance.

$$T_s = 0.5\,(T_o - S\sqrt{L_s}) \qquad \textbf{(3A)}$$

where $S = (T_o - T_r)/\sqrt{L_r}$ **(3B)**
and T_s = number of turns to remove from each winding to obtain L_s
L_s = desired inductance (mH)
T_o = previously determined number of turns on the unmodified inductor
T_r = total turns removed to get L_r
L_r = inductance (mH) when T_r turns are removed

Example: a 134-mH inductor has a $T_o = 1064$, with $L_r = 110$ for $T_r = 100$

$$S = (1064 - 100)/\sqrt{110} = 91.914$$

then $T_s = 0.5\,(1064 - 91.914\sqrt{L_s})$

$$= 532 - 45.957\sqrt{L_s}$$

For example, find T_s to get 132 mH from the 134-mH inductor.

$$T_s = 532 - 45.957\sqrt{132} = 532 - 528.0$$

= *4 turns removed from each winding.*
Total turns removed = 8

For a parallel-aiding connection:

$$T_p = T_o/2 - S\sqrt{L_p} \qquad \textbf{(4)}$$

where T_o and S were previously defined, and
T_p = number of turns to remove from each winding to obtain L_p
L_p = desired inductance (mH)

$$T_p = 91.914\sqrt{L_p} \qquad \textbf{(5)}$$

Find T_p in order to obtain 33 mH from the 134-mH inductor with its windings in the parallel-aiding connection:

$T_p = 532 - 91.914\sqrt{33} = 532\text{-}528.0$
= *4 turns removed from each winding*
Total turns removed = 8

appendix B

Equations relating reflection coefficient and A_p:

$$R.C.\ (percent) = 100\sqrt{1 - 0.1^x} \qquad \textbf{(1)}$$

where $x = A_p/10$ and A_p is the maximum passband ripple amplitude in dB.

$$A_p(dB) = -10 \cdot \log_{10}(1 - \varrho^2) \qquad \textbf{(2)}$$

where ϱ = *percent R.C./100*

For example, if $R.C.$ = *8.058 percent*, then A_p = *0.0282913 dB.*

appendix C

determination of filter matching pad resistances

When the source impedance (Z_s) or load impedance (Z_L) is within 30 percent of the termination impedance required by the filter, a single resistor can be placed in series or parallel with Z_s or Z_L to produce the termination impedance required by the filter. The slight loss in signal level can usually be compensated for by simply increasing the volume control on your receiver. For greater impedance differences, a suitable matching transformer must be used (see note 1).

(A) When Z_s or Z_L *is less than* the required filter termination impedance place a resistance *in series* with the Z_s or Z_L so that the sum of the two equals the desired filter termination impedance.

(B) When Z_s or Z_L *is greater than* the required filter termination impedance place a resistance *in parallel* with the Z_s or Z_L so that the parallel combination equals the desired filter termination impedance. If the added parallel resistance is "R_p," then:

$$R_p = (S \cdot F)/(S - F)$$

where S = source impedance and
F = filter impedance with all values in ohms.

For example, if a 500-ohm filter is to be matched to a 350-ohm source and a 600-ohm load, then place a 150-ohm resistor in series with the 350-ohm source, and connect a 3 kilohm resistor in parallel with the 600-ohm load.

Note 1: A large selection of inexpensive audio transformers of different impedances at power levels of 0.075, 0.2, 0.4, and 2-watts output can be obtained from Mouser Electronics, 11433 Woodside Avenue, Santee, California 92071 (714-449-2222). Prices vary from 79 cents to $1.20 for the 0.075 to 0.4-watt levels and $2.30 for the 2-watt level.

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ham radio

high frequency receiver performance

Understanding and specifying good receiver operation

Receiver performance and the means for its specification is a complex and frequently misunderstood topic, for it involves many different and often unrelated parameters, all of which are important to the overall performance of the receiver. (For example, it's no good having exceptional sensitivity if the receiver drifts excessively.) When specifying performance, it is important that all parameters be specified accurately and completely. Many manufacturers and reviewers have been guilty of specifying performance inadequately, with ambiguous and misleading results. For example, the frequently quoted "sensitivity 0.5 μV" is meaningless, and "0.5 μV for 10 dB" is not much better.

It is interesting to note that the importance of some performance parameters (e.g., intermodulation and reciprocal mixing) have only recently been fully appreciated and understood. Also certain receiver design changes have actually been responsible for *reducing* performance: for example, the direct replacement of tubes by bipolar transistors in the late 1960's, and the current widespread use of frequency synthesis, which can introduce spurious receiver responses if not designed with special care.

noise and sensitivity

One of the fundamental concepts underlying receiver performance is that of noise. So before we examine performance parameters — what they mean, how they are specified, and how they can be improved — we'll take a brief look at the physics of noise.

Thermal noise is due to the random movement of particles in the effective impedance at the input to the receiver at ambient temperature. It cannot be avoided (except by cooling the whole antenna system down to near absolute zero) and is even present if a shielded 50 or 75 ohm resistor representing the antenna impedance is plugged into the receiver input. It is given by:

$$V = \sqrt{4kTBR} \qquad (1)$$

where:
V = RMS noise voltage
k = Boltzmann's constant
T = temperature in degrees Kelvin
B = receiver bandwidth in Hz
R = resistance/impedance in ohms

For a given antenna impedance and a nominal ambient temperature (taken as 300 degrees K or 27

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degrees C), the only way to reduce the thermal noise is to reduce the receiver bandwidth. For a 50 ohm impedance, ambient temperature of 300 degrees K, and a bandwidth of 3 kHz the above expression works out as:

$$V = \sqrt{(4)(1.38 \times 10^{-23})(300)(3000)(50)}$$
$$= 0.05\,\mu V\,(EMF) = -26\,dB\mu V$$

where a dBμV is a dB relative to 1 microvolt *(EMF)*.

Sensitivity is expressed as the smallest signal required to give a specific signal-to-noise ratio (S/N),* in a particular receiver bandwidth, and in the case of AM, for a given modulation level. In the days of noisy tubes, good sensitivity was hard to achieve without compromising other performance aspects. Now with bipolar transistors and FETs, sensitivities of 0.5 μV for a 10 dB S/N ratio in a 3 kHz bandwidth, can easily be achieved on HF for an SSB or CW signal. As AM is usually specified at 30 percent modulation level, the figure will be 10 dB (3.16 times) worse than the SSB/CW figure; in this case, 1.6 μV (EMF).

Noise factor (NF) can be defined as the ratio of the S/N of a hypothetically perfect (noiseless) receiver to that of a real receiver, which adds its own noise to the thermal noise. Since it is the ratio of two ratios it is independent of bandwidth, temperature, and impedance.

Typical HF spectrum noise voltages within a 3 kHz bandwidth are shown in **fig. 1**. We see at -26 dBμV the thermal threshold noise, and also the receiver noise for a NF of 10 dB, at -16 dBμV. To achieve a S/N of 10 dB a signal will need to be at -6 dBμV or 0.5 μV, which means that an NF of 10 dB is equivalent to a sensitivity of approximately 0.5 μV for a 10 dB S/N in a 3 kHz bandwidth.

However, the most significant conclusions to be drawn from **fig. 1** are concerned with atmospheric noise. This is plotted on **fig. 1** for a quiet area at a

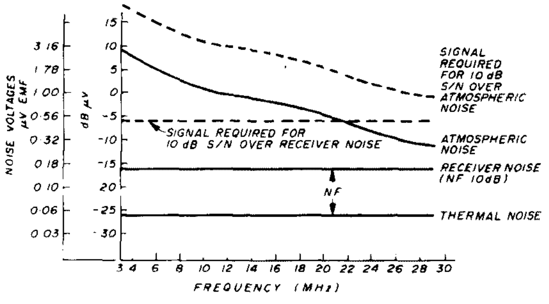

fig. 1. HF spectrum noise in a 3 kHz bandwidth.

*In fact it's more appropriate to quote signal + noise to noise (S + N/N) or even signal + noise + distortion to noise + distortion (SINAD). Note for an S/N of 10 dB or more there is little difference between the three terms.

quiet time, and is between 5 and 25 dB above receiver noise; consequently for a receiver with a noise figure of 10 dB it is the atmospheric noise, not receiver noise, that limits receiver performance. Indeed the NF could be increased to 15 dB (1μV EMF sensitivity) without loss of performance, except perhaps at 25 to 30 MHz. There is, therefore, no point in reducing the NF below 10 dB — especially as sensitivity can only be obtained at the expense of dynamic effects such as intermodulation. It is also worth noting that claims of 0.15 μV (EMF) for 10 dB S/N (seen recently for an SSB transceiver) are quite impossible. Even a perfect receiver with a 0 dB NF needs 0.16 μV (-16 dBμV) to achieve 10 dB S/N due to the thermal threshold of -26 dBμV.

From **fig. 1** it can be seen that under real operating conditions a receiver with a sensitivity of 0.5 μV will in fact need between 1 μV (at 30 MHz) and 10 μV (3 MHz) for a 10 dB S/N ratio on HF — and this for a quiet atmosphere (and no QRM)!

The above discussion considered noise voltage in a 3 kHz receiver bandwidth. However, noise is proportional to the square root of the bandwidth. Thus if bandwidth is reduced from 3 kHz to 300 Hz, all noise voltages (thermal, receiver, and atmospheric) drop by a factor of $\sqrt{10}$ or 3.16 times (10 dB). This explains the continuing use of CW in the HF bands; a CW signal can still be copied when SSB would be lost in the noise. In fact, some operators can copy a CW signal with a S/N of around 0 dB, so the advantage over SSB can be as much as 20 dB†. Thus sensitivity for the same (10 dB NF) receiver can be quoted as 0.05 μV for 0 dB S/N in a 300 Hz bandwidth.

Note that all voltages above are EMF (Electromotive Force). Recently it has become common to specify sensitivities and other parameters using potential difference (PD) instead. This practice makes sensitivity figures look twice as good (6 dB better) because, in a matched impedance system, PD is *always* half EMF (see Appendix). Care must therefore be exercised when interpreting receiver specifications; often the distinction is not made clear, usually indicating that PD is intended (see **dynamic range** for explanation).

selectivity

Selectivity used to be achieved by means of distributed tuned circuits in the IF strip, and to obtain good selectivity a low second IF was required (for example, 455 kHz). Selectivity is now usually obtained by means of crystal, mechanical, or ceramic block filters. The ideal filter response has a flat top with low ripple, and steep sides going down to a -70 dB (or

†If one considers that the human ear consists of a contiguous series of extremely narrow high Q filters (less than 30 Hz) then perhaps the ear/brain combination always works with a positive S/N ratio for intelligibility. **Editor**

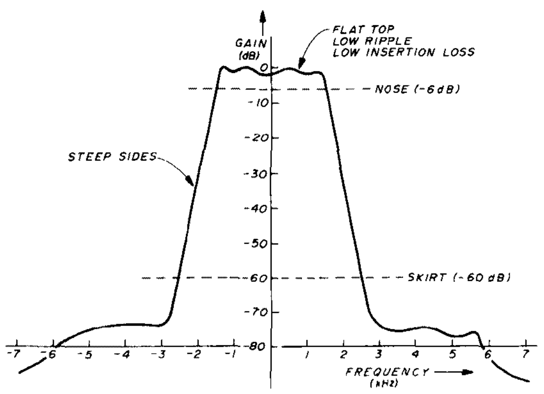

fig. 2. A desirable filter response would include a fast roll-off, low passband ripple and insertion loss, and a high stopband rejection.

greater) stopband, and extending a long way out (see **fig. 2**).

The old constraint of a low second IF no longer applies. In fact, it is easier to design crystal filters for higher frequencies; IFs of 1.4, 1.6, 9 and 10.7 MHz often being used.

Selectivity is usually quoted at the nose bandwidth (6 dB down), and the skirt bandwidth at 60 dB down. Good values for an 8 pole SSB filter are 2.7 kHz and 4.4 kHz, respectively. One measure of filter performance often quoted is the **Shape Factor** (SF), which is the ratio of the skirt bandwidth to the nose bandwidth. The ideal SF is 1:1, and anything less than 2:1 for a 3 kHz SSB filter is considered good.

Mechanical and crystal filters can get quite close to the ideal, and some less expensive ceramic filters give surprisingly good results. Impedance matching into and out of a filter is of great importance and insertion loss (the loss caused by the filter in the middle of the passband) must be made up by amplification (usually less than 10 dB). A typical set of filters for a high grade communications receiver might be 8 kHz (AM), 2.7 kHz (SSB) — often with two asymmetrical filters, one for USB one for LSB — and 1.0 kHz, 300 Hz, and 100 Hz for CW. The trend in Amateur equipment is for the tightest possible SSB filter (2.4 kHz), and often 600 Hz or 300 Hz for CW are used.

The above refers to what could be called the "static" selectivity of the receiver, or the selectivity to a single signal only. For a discussion on dynamic selectivity see the section on **reciprocal mixing**.

As an interesting aside, consider the use of the audio CW filter, often used by the Amateur fraternity in lieu of a good CW filter at the IF. (The best filter of all, of course, is the human brain; a good operator can pick out and copy a weak CW signal in company with numerous other signals because of the difference in tone. Many experienced CW operators prefer to listen in a wide bandwidth and do their *own* filtering even when sharp CW filters are available.) However, the audio image frequency (see **fig. 3**) will also give an output of exactly the same tone, and even the "human filter" will find it impossible to differentiate. But this is exactly the frequency that the audio filter cannot differentiate, either! Also, unless AGC voltage is at least partially audio-derived, strong unwanted signals in the IF passband will reduce the IF gain, hence reducing post-filter dynamic range.

Nevertheless a good multi-pole audio filter can be beneficial, especially if a linear detector is being used. (A product detector is linear; an envelope detector is not.) Also, if a steep-sided SSB filter is available, the audio image can be rejected if the BFO injection is made to coincide with the edge of the passband.

Image (second channel) rejection. In the normal superheterodyning process, a wanted signal (F_S) beats in the mixer with the local oscillator (or synthesizer output) frequency (F_{LO}), and one of the resultant products of the mixing process, usually

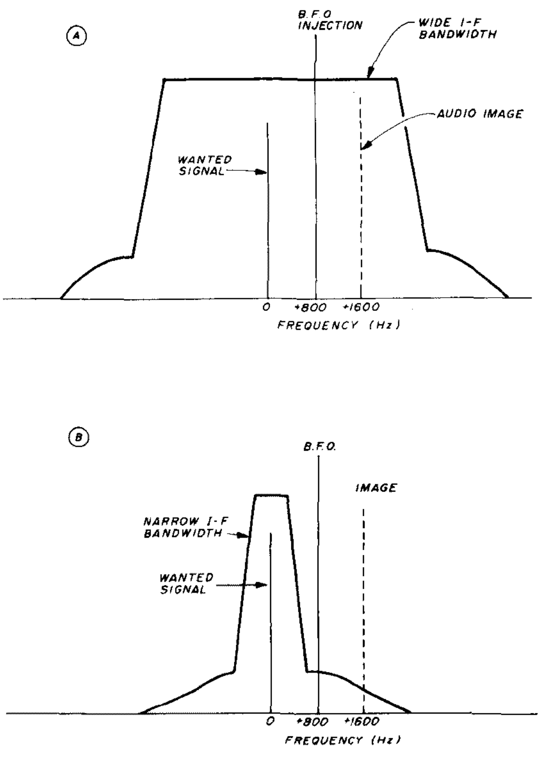

fig. 3. Narrowing the IF bandwidth improves the audio-image rejection: (A) wide I-F bandwidth; (B) narrow I-F bandwidth.

$F_{LO} - F_S$, at the intermediate frequency (IF), is passed by the IF selectivity filter.

However, another frequency called the image or second channel frequency ($F_{LO} + F_{IF}$) also beats with the local oscillator frequency to produce a product at the IF. This frequency must be rejected by some form of RF tuning, either ganged to the "tune" control or using a separate "pre-select" control; or by means of switched bandpass filters, usually automatically switched on synthesized receivers.

As the image frequency is equal to F_S plus twice the IF, the higher the first IF, the further away from F_S will be the image frequency, and the easier it will be to reject. If up-conversion techniques are used, the first IF will be in the range 40 to 90 MHz and the image frequency will also be at VHF and thus can be rejected by a simple 35 MHz low-pass filter at the receiver input.[3]

Image frequency rejection is specified as the ratio in dB of an unwanted signal above 1 μV to give the same output as the received 1 μV signal. 50 dB of rejection is considered poor, while 80 dB or more is desirable.

IF rejection

IF rejection occurs when a strong signal *at a receiver IF* directly breaks through the early stages and into the IF. It is specified in the same manner as image rejection with 80 dB being a desirable number. It should be quoted for all the IFs in a receiver. Often in a double conversion receiver the figure for the second IF is worse than that for the first IF. Good screening is necessary between IF and RF stages, and IF traps in early pre-IF stages can be employed to reduce IF breakthrough. Taken together, IF and image rejection are sometimes referred to as "rejection to external spurii."

dynamic effects

Dynamic interference effects such as intermodulation and cross-modulation have often been largely ignored in the past, and only in the last ten years or so has their true importance been understood. It has to be said that the replacement of tubes by bipolar transistors in the 1960s made the situation worse.* In general, dynamic effects are caused by large off-tune (off frequency) signals that cause the receiver to operate in a non-linear manner. It is these effects (together with reciprocal mixing) that currently determine the performance of the communications receiver rather than the traditional parameters of sensitivity, selectivity, and stability.

dynamic range and intercept point

Dynamic range can be loosely described as the range of input signals over which dynamic interference effects produce outputs which are not significant (i.e., which are below the noise floor). In order to arrive at a suitable definition for dynamic range, consider the intercept point as shown in **fig. 4**. This occurs because the power level of the dynamic product increases at a greater rate than that for the wanted signal. Second order products increase as the square of the input (twice as many dB), and third order products as the cube (3 times in dB). (Third and second order products will be dealt with in greater detail when discussing intermodulation.)

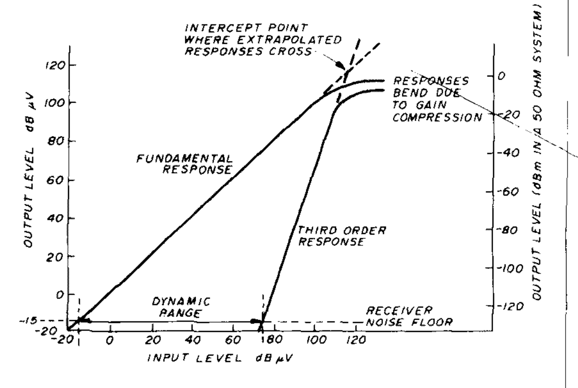

fig. 4. Intercept point is determined from the intersection of the fundamental and third-order response curves.

Dynamic range can then be usefully defined as two-thirds of *the difference in level between noise* floor and the intercept point, or alternatively as the difference between the fundamental response input level and the third order response input level as measured along the noise floor (see **fig. 4**). These are by no means the only methods of specifying dynamic range, and care must be taken in interpreting manufacturers figures. Using the above method of definition a dynamic range of 90 to 100 dB for 3 kHz bandwidth with an intercept point of 120 to 140 dBμV (+7 to +27 dBm), can be considered good. Note that because dynamic range — by definition — depends on the noise floor level, it will increase as bandwidth is reduced. Intercept point is sometimes specified (as are other parameters) in dBm, where a dBm is a dB relative to 1 mW into the system impedance (usually 50 ohms). 0 dBm (50 ohms) equals 224 mV PD (Potential Difference), or 113 dBμV, thus to convert from dBμV to dBm (50 ohms) simply subtract 113. (E.g., *0 dBμV = 1 μV EMF = 0.5 μV PD is −113 dBm).* Note: dBμV is dB relative 1 μV EMF.

*Readers are urged to consider the outstanding intermod performance achieved by using FETs these days. **Editor**

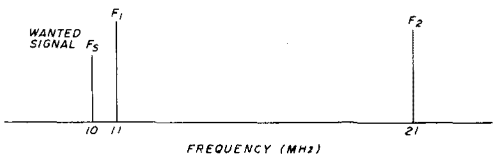

fig. 5. Example of second-order IMDs.

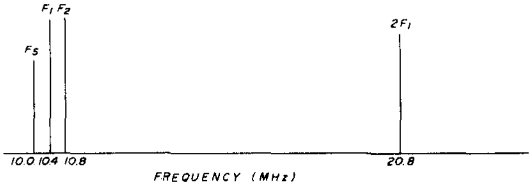

fig. 6. Example of third-order IMDs where $2F_1$ beats with F_2 to produce a response at 10.0 MHz.

intermodulation

Intermodulation product distortion (IMD) occurs when two large unwanted signals beat together (inter-modulate) in a non-linear receiver stage, to produce a product at the *wanted* frequency.

Second order intermodulation products equate to $F_1 \pm F_2$ (where F_1 and F_2 are the two unwanted frequencies). This process is shown in **fig. 5** where the two unwanted signals are at 11 and 21 MHz causing a beat at 10 MHz. (Signals at 6 and 16 MHz or 3 and 7 MHz would produce a similar product.)

One point to note about second order IMDs is that at least one of the unwanted signals must be outside the passband of any octave filter response that includes F_S and consequently can be rejected by any reasonably tight RF tuning, such as a good octave filter or sub-octave (less than an octave) filter.

Third-order intermodulation products are equal in frequency to $F_1 \pm 2F_2$. Thus both signals can be very close to the wanted signal, and well within the RF passband regardless of the type of RF tuning in use. This is seen in **fig. 6** where signals at 10.4 and 10.8 MHz produce a third-order IMD product of 10 MHz.

Intermodulation performance is typically specified as the levels of two unwanted signals not less than 20 kHz off tune (off the main signal frequency) to give a 0 dBμV (1 μV EMF) response. A good receiver will have a third-order IMD performance of 70 to 90 dBμV. A statistical analysis based on data relating to actual signals received over the whole of the HF band on wideband (rhombic) antennas[1,2] indicates that at least 90 dB is required. As specified above, this corresponds to a dynamic range of 100 dB and an intercept point of 134 dBμV or +21 dBm (50 ohms), as shown in **fig. 7**. It is easy to see why such a figure is needed. 90 dBμV corresponds to 32 mV EMF and at almost any time there will be tens of broadcast (and other) stations putting between 10 and 100 mV onto a wideband antenna, with hundreds of others in the range 1 to 10 mV (EMF)!

Second-order IMD performance, often not stated, can sometimes be misleading. Second-order performance can be poor enough to present a real problem even with good third-order performance if a wideband antenna is used without an antenna tuning unit (ATU). This is due to the use of unshielded octal filters of less than perfect performance. Levels similar to third-order performance are *required.*

In-band intermodulation occurs when two signals *within* the IF passband beat (mix) to produce extra products. It is normally of little significance in HF communications except where multichannel "Voice Frequency Telegraphy (VFT)" systems such as "Piccolo" are in use. A typical level of performance for a good receiver is for a product of −40 dB with reference to two in-bands signals.

Cross modulation occurs when modulation from an unwanted signal transfers itself across and "modulates" the wanted signal. Again this is due to non-linearities in the early receiver stages, and sometimes the same modulation will reappear on each adjacent signal tuned in. Cross modulation is a third-order effect, so good third-order IMD performance will tend to mean good cross modulation performance (see **fig. 8**).

Cross modulation may be specified as the level required in dBμV for a 30 percent modulated carrier greater than 20 kHz off channel to cause 3 percent cross modulation. A level of 70 to 90 dBμV is considered good.

Blocking, or de-sensitizing, is similar to cross modulation, but in this case the large off-channel signal causes a reduction in wanted signal output. It is specified as the signal required to reduce wanted output by 1 dB. It can often be caused by a strong CW signal, causing gain to go up and down with the keying. 90 to 110 dBμV is considered good performance, for a 1 mV (EMF) signal.

causes and cures of dynamic effects

Dynamic effects are caused by large off-channel signals driving the receiver into non-linearity. There are three fundamental methods of improving performance: (*a*) preventing the off-channel signals

a performance specification

This detailed specification indicates the performance required for a "very good" receiver. Additional parameters not mentioned previously, such as audio output power, have been added for completeness. Note that specifications like this should state worst-case figures (maximum or minimum values that are acceptable when the receiver is being tested); in some cases, however, "typical" values are used instead. This will at first sight make the receiver seem considerably better: for example, a worst-case NF of 10 dB will often (typically) be 7 dB. It is therefore important to establish whether a true specification or a list of typical values is meant. The specification below usss worst-case figures. Note dBμV is dB relative to 1 μV EMF.

specification

frequency coverage	60 kHz to 30 MHz (continuous coverage, using full synthesis)
frequency display	7 digit, resolution 10 Hz
reception modes	CW, FSK (RTTY), SSB (USB and LSB), ISB, AM, NBFM
input impedance	50 ohms
sensitivity (600 kHz to 30 MHz)	SSB, CW, ISB - 0.5 μV EMF for 10 dB SINAD in 3 kHz AM - 2.2 μV EMF for 10 dB SINAD in 8 kHz, modulation 30 percent FM - 0.8 μV EMF for 10 dB quieting in 15 kHz bandwidth, 60 percent deviation (Noise factor 10 dB)
IF selectivity	

filter	available on	response (kHz) −6 dB	response (kHz) −60 dB	shape factor
15 kHz	FM	15.00	33.00	2.2
8 kHz	AM	7.80	14.00	1.8
2.7 kHz	SSB, ISB, CW	2.70	4.30	1.8
1 kHz	CW	1.00	2.00	2.0
300 Hz	CW	0.28	0.75	2.7

dynamic range	104 dB. (Intercept point 140 dBμV = +27 dBm (50 ohms)) (3 kHz bandwidth).
third-order intermodulation	94 dBμV. The levels of two unwanted signals both greater than 20 kHz off frequency will be 94 dBμV (or −18 dBm (50 ohms)) to give a 1 μV (0 dBμV) response.
second-order intermodulation	90 dBμV. The levels of two unwanted signals both greater than 20 kHz off frequency will be 90 dBμV (−23 dBm (50 ohms)) to give a 1 μV (0 dBμV) response.
in-band intermodulation	−40 dB. Two signals in-band of equal amplitude will produce a product greater than 40 dB down.
cross modulation	100 dBμV. A 30 percent modulated carrier greater than 20 kHz off frequency must be 100 dBμV (−13 dBm (50 ohms)) to cause 3 percent cross modulation.
blocking	100 dBμV. A signal greater than 20 kHz off frequency will be 110 dBμV (−3 dBm (50 ohms)) to cause a 3 dB reduction of wanted 1 mV EMF (60 dBμV) signal.
reciprocal mixing	90 dB. An unwanted signal 50 kHz off frequency will be 90 dB above the level of a wanted on-frequency signal to reduce its SINAD by 3 dB, in a 3 kHz bandwidth.
image rejection	90 dBμV. The image frequency must be 90 dBμV (−23 dBm (50 ohms)) to give a 1 μV (0 dBμV) response.
IF rejection	90 dBμV. For the first and second IF the level of an unwanted signal will be 90 dBμV (−23 dBm (50 ohms)) to give a 1 μV (0 dBμ) response.
crosstalk	−50 dB. On ISB mode the crosstalk between two equal 0 dBm (600 ohm) outputs shall be less than −50 dB, relative to output at 1 kHz.
response to internal spurii	All internally generated spurious signals are less than 3 dB above receiver noise. (3 kHz bandwidth.)
stability	After 10 minute warmup, better than 1 part in 10^8/°C. Long term crystal aging less than 1 part in 10^9 per day.
antenna radiation	Less than 10 μV_{PD} (20 μV EMF) into 50 ohms (−87 dBm). (2pW)
protection	Receiver can withstand 30 volts at antenna input continuously. A spark gap is provided.
AGC performance	For an input change of 90 dB, the output change will be less than 3 dB for all signal levels greater than 3 μV_{EMF} (10 dBμV).
AGC time constants	

	SLOW attack	SLOW hang	SLOW decay	FAST attack	FAST hang	FAST decay
ISB, SSB, CW	10 ms	2s	200 ms	5 ms	200 ms	60 ms
AM, FM	20 ms	—	20 ms	5 ms	—	5 ms

BFO range (CW)	±3 kHz
AF outputs	Main output: 2W into 8 ohms at less than 3 percent Total Harmonic Distortion (THD) Headphones: 20 mW into 600 ohms Line: 600 ohms balanced line independent of AF gain control, settable to −10 to +10 dBm (600 ohms), i.e. 245 mV to 2.5 VPD.
muting	0V on mute terminal for muting. Mute level at least 60 dB down.
metering	Meter reads signal strength in dBμV or AF line level in dBm (600 ohms).
power requirements	200 to 250 VAC, 45 to 65 Hz, at 50 VA (Volt-Amperes)
environment	Operating: 14 to 104°F (−10 to +40°C) Storage: −40 to 158°F (−40 to +70°C) Relative humidity up to 95 percent at 104°F (40°C) (non-condensing).
MTBF (mean time between failure)	8000 hours

from entering; (*b*) improving the linearity of the early stages of the receiver, prior to and including the roofing filter (a roofing filter is inserted in the first IF and is used to reduce the number of strong signals passing through the IF [chain]; it is not as narrow as the main selective filters); and (*c*) reducing the level of all signals.

This last method works because the response to the unwanted (dynamic) signals falls off at a faster rate than that of the wanted signals (see **fig. 4**). It is implemented by means of a front-end attenuator or by a wideband AGC loop (separate from the main AGC loop) which operates on the RF amplifier on large signals only, and thus can be thought of as being an automatic attenuator. Both methods have the disadvantage of reducing receiver sensitivity, and consequently other solutions should be found.

Method *(a)* involves the use of sub-octave filters or some sort of preselector tuning and can be very effective in reducing second order effects. However, as previously mentioned, third order products can be too close (in the signal passband) for tuned circuits to have an effect.

The only real solution is to improve linearity, (*b*). Bipolar transistors are particularly poor in this respect, but FETs are approximately square-law devices and are therefore very good in terms of third-order effects, but not as good for handling second-order products. Linearity can be improved by using higher power supply voltages and by keeping pre-roofing filter gain down to a minimum consistent with required sensitivity, and therefore by keeping

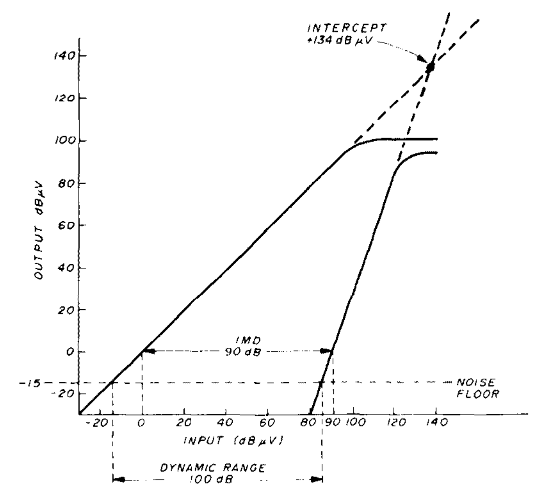

fig. 7. Good receiver third-order intermodulation distortion performance is shown as a 100 dB spurious-free dynamic range.

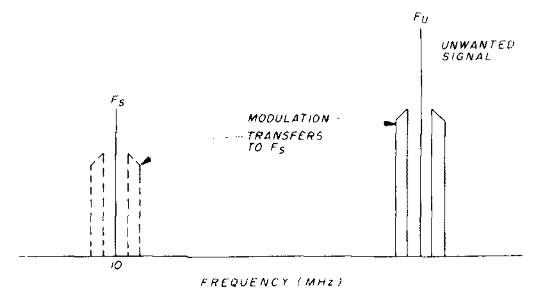

fig. 8. Cross modulation occurs when modulation from an "unwanted" signal transfers itself across and modulates the wanted signal.

noise levels down. In the extreme situation, the RF amplifier can be completely eliminated and the signals fed directly to a low-noise mixer via the front-end filters. This isn't such a drastic step as it might appear, because remember an NF of 15 dB is more than adequate on HF, and this can be achieved without an RF amplifier.

The mixer may be a double-balanced switching type diode mixer using high level LO injection to improve linearity. Components normally considered to be linear, passive, and reciprocal must be carefully checked to ensure that they are. This especially applies to ferrite cores used for RF coils and transformers; and crystal filters, which are often non-linear and non-reciprocal, i.e. having different characteristics if connected the "wrong" way around.

The practice of fitting protection diodes at the receiver input (often found on marine-band receivers) will also cause non-linearity, as will diodes used to switch filters. If all these points are carefully considered, very good linearity can be achieved with an intercept point of 140 dBμV or better. This level of performance ensures that IMD cross-modulation products are below atmospheric noise on HF.

Reciprocal mixing is due to high level unwanted signals mixing with the noise sidebands of the local oscillator/synthesizer, producing noise products at the receive frequency (see **fig. 9**).

It is another phenomenon which until recently has been more or less ignored, partly because tube local oscillators are inherently "cleaner" than most of the modern solid-state synthesizers. It can be specified as the level in dB above a wanted signal that an unwanted signal can attain at a specified frequency off-channel (e.g. 50 kHz) at a specified bandwidth (usually 3 kHz) to reduce the S/N ratio of a wanted on-channel signal by 3 dB!

The oscillator noise can be reduced by employing

high "Q" (oscillator) circuits, and also by using high power oscillators, as the noise sidebands will then be relatively weaker, thus improving S/N. Phase locked loops can be very poor in this respect because they contain numerous noise sources which add together in the output, together with frequency jitter. In addition, PLL's frequently use low "Q" circuits and low power VCOs in the output. The noise produced is phase modulated and cannot be removed by limiting.

The action of reciprocal mixing in introducing off-channel signals into the IF at levels proportional to the distance away from the wanted signal (frequency separation) (see **fig. 9**) effectively reduces the selectivity of the receiver. This is shown in **fig. 10**, and the response curves indicate the dynamic selectivity of the receiver, that is, the selectivity of the receiver in a real signal environment. As can be seen, it's the stopband of the filter response that's been changed; with 70 dB reciprocal mixing, a considerable loss of performance occurs.

However, when the reciprocal mixing has been improved to 90 dB, its effect on filter response can be considered fairly minor. A frequency-synthesized receiver can achieve 90 dB, while a good (tube or FET) crystal oscillator can give 110 dB or better.

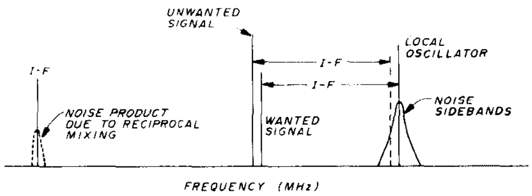

fig. 9. Reciprocal mixing reduces the useful dynamic range of the receiver.

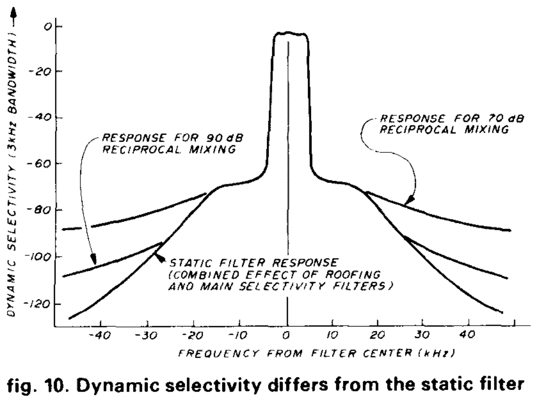

fig. 10. Dynamic selectivity differs from the static filter response because of reciprocal mixing.

synthesizer noise and internal spurious responses

Internal spurious responses are responses of the receiver to signals generated within the receiver itself. These internally generated signals can be fixed (e.g., reference frequencies) or can move when the receiver tuning is changed. They cause problems when they occur at the signal frequency or an IF, and are generated by any oscillators within the receiver or by digital circuitry such as synthesizers and frequency counters. (A current trend is to use multiplexed fluorescent displays instead of DC driven LED displays, which generate low frequency interference. Adequate shielding must be employed to reduce radiation.) In using frequency synthesis techniques, many of the waveforms are digital, square waves with fast rise-times rich in harmonics. CMOS and LSI (N-MOS) is usually better in this respect than TTL which has faster rise times. Careful design, with adequate low-pass and bandpass filtering and with high "Q" output circuits, is important. With good design it is possible to keep spurious outputs 100 dB down on the main output level. This standard of performance should ensure that all spurious responses are no more than 3 dB above the receiver noise level.

stability

A fully synthesized receiver can have a stability equal to that of the frequency reference source.[3] If an oven-temperature stabilized crystal oscillator is used, stability of less than 0.1 Hz/°C can be achieved. With partial synthesis the stability is governed by the stability of the VFO, but with cool, buffered solid-state designs it is possible to achieve long term drift rates of 100 Hz/hour with short term drift (even including lifting the receiver an inch and dropping it!) of 10 to 20 Hz. The latter is more than adequate for normal SSB/CW/RTTY/AM communications.

conclusions

The performance quality of the HF receiver has increased over the years and it is now possible to design a general coverage receiver that provides very high performance. As activity on the HF bands has constantly increased, this improvement in performance is of vital importance to maintain the ability to communicate.

Generally speaking, the cost of equipment has gone up in proportion to its complexity. There is, however, a reduction in real cost due, in part, to the availability of relatively low cost complex ICs, crystal filters, and FETs, as well as to improvements in design techniques.

Future trends will include more extensive use of

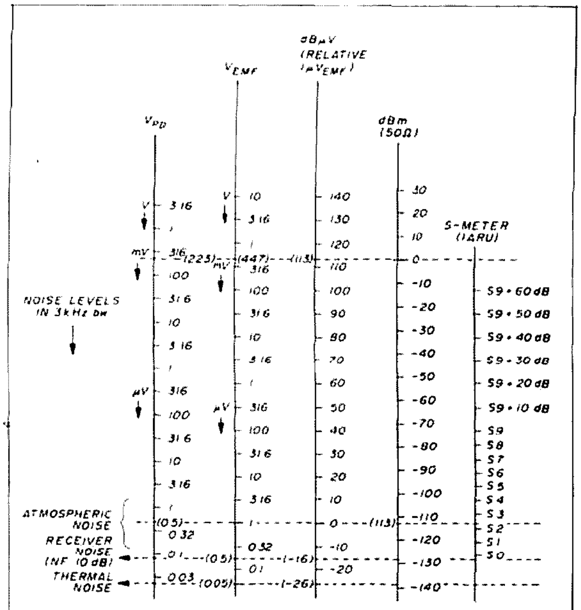

fig. 11. Voltage and power relationships.

ICs, with possibly a single chip full-synthesizer, and microprocessor control of the receiver. It seems likely that more extensive use will be made of remote control of receivers via data link, and that in communications centers a central mini-computer will be linked to each operating position, performing a variety of useful functions. On the domestic scene, the home computer can be linked to the receiver via an RS232 line, and can be used to decode RTTY and SSTV signals, etc. It could also be used as a big "memory" to store channels (frequencies/modes/filters and a channel ident) for instant recall.

In conclusion, receivers available now offer performance that ten years ago could be obtained only from professional receivers costing ten times as much, and which twenty years ago could not be obtained at all.

appendix

Fig. 11 shows the relationships between levels as specified in V(EMF), V(PD), dBμV, and dBm (50 ohms). Also shown is the S-meter response as recommended by the IARU, the International Amateur Radio Union. This specifies that S9 should be at 50 μV PD, or 100 μV EMF; and that each S-point should be at 6 dB intervals. Also shown are the thermal noise threshold and typical receiver and atmospheric noise terms for a 3 kHz bandwidth.

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ham radio

VHF/UHF WORLD

Joe Reisert W1JR

VHF/UHF antennas and antenna systems

There is probably no other VHF/UHF topic that inspires such a vigorous line of conversation than the subject of antennas. Hardly a day goes by when I'm not asked questions like "What type of antenna do you recommend?" or "What's the difference between this antenna and that one?" The reason we're so interested in this subject is that there is no better way to improve the performance of your station: *a 1 dB improvement in antenna gain yields a 2 dB overall increase in your station capability*, since we gain 1 dB on transmit and 1 dB on receive.

The answers to the questions above are neither simple nor obvious; they take time to answer thoroughly. Furthermore, the answers may not always involve performance as much as economic or structural considerations. With this in mind, I've decided to discuss the popular types of VHF/UHF antennas and the tradeoffs between them in this month's column. Future columns will zero in on ways to measure and obtain peak performance with the antennas we are using.

antenna types

The three major types of VHF/UHF antennas presently in use (excluding FM/repeater types) are the collinear array, the Yagi and the parabolic dish. Each type has its own advantages and disadvantages. While all can produce high gain, they differ vastly in form factor. Let's discuss each type individually and see what each can and can't do.

collinear array

The collinear array was very popular among VHF/UHF'ers and especially EME'ers (Earth-Moon-Earth) on 2 meters and 70 cm (432 MHz) before high performance VHF/UHF Yagi antennas were designed. It usually consists of a group of half-wavelength dipoles in front of a screen or set of reflectors (**fig. 1**). Technically speaking, you could probably call this an array of 2-element Yagi antennas. Some individuals have even placed additional directors in front of the driven dipoles in an attempt to increase gain.[1] The unique thing about a collinear array is that the feed system is usually some form of open wire line. The collinear, unlike many antennas, is usually quite broadband. Efficiency can be very high and gain is mainly a function of the size of the array. The extended-expanded collinear is a stretched-out version that has fewer elements and higher gain approaching 80 to 95 percent efficiency.[2] This antenna is treated in depth in reference 2 and improvements are mentioned in reference 3.

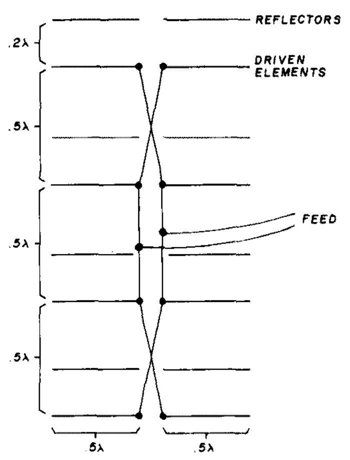

fig. 1. Typical collinear array.

Collinear antennas offer several advantages: They are not "critical" to build, are usually low in cost, have a low loss feed system, and when used on EME, can be readily adapted to polarity rotation. They are also easy to array when extremely high gain is desired. The principal drawback is their overall size, which usually precludes mounting other antennas on the same mast. (I'll discuss this subject later in this column.) There may also be problems in areas where moisture is present, since the VSWR may increase if the feed system gets damp or wet.

Yagi and Yagi type antennas

The Yagi antenna (**fig. 2**) is particularly popular where space is at a premium or when only narrowband operation is required. It is presently the workhorse in the VHF and lower UHF spectrum. The first high gain VHF Yagi designs were published by Kmosko and Johnson,[4] Greenblum,[5] and Ehrenspeck and Poehler.[6] Unfortunately, these Yagis weren't always

as good as claimed, were hard to duplicate, and when duplicated, often failed to deliver the promised gain. Some had very high side lobes.

The WØEYE 70 cm 4.2 wavelength 15 element Yagi[7] with corrections[8] was first published in 1972. Based on the unpublished work of Pete Viezbicke at NBS (National Bureau of Standards), it was the first really high gain Yagi with a clean pattern that was easily duplicated. Don Hilliard, WØEYE (now WØPW), and I urged Pete to publish his work and he finally did so in December, 1977, in *NBS Technical Note 688*, now out of print.[9] This publication, the result of extensive studies done by NBS in the 1950's to develop high gain arrays for ionospheric scatter, included six different models with boomlengths of 0.4 to 4.2 wavelengths. In the August, 1977 issue of *ham radio*, I published an article which included the

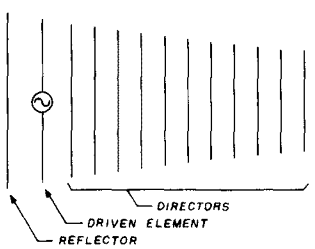

fig. 2. Typical long-Yagi.

majority of the NBS material along with several practical examples of Yagi designs using this material.[10] (A future column will deal with this topic in greater detail.)

The NBS Yagi designs described above work well through 70 cm, but the longest one is only 4.2 wavelengths (approximately 10 feet or 3 meters at 70 cm). Günter Hoch, DL6WU, has been working on this problem and has recently produced some very long Yagis[11] based on an extension of the work of Greenblum.[5] Hoch has been able to increase the gain approximately 2.35 dB each time the boomlength is doubled and has demonstrated this success even at 23 cm (1296 MHz). I have verified his information with a 70 cm 9.25 wavelength (21 feet or 6.5 meters) model which measured greater than 17.5 dB over a dipole gain at the 1981 Central States VHF Conference.[12] At the 1983 Central States VHF Conference, KL7WE entered an even longer (24 feet or 7.5 meter) 70 cm Yagi model based on the same information and it measured approximately 18 dB over a dipole. Hence the quest for extremely high gain Yagis is finally showing promise.

the Quagi

In April 1977 Wayne Overbeck, K6YNB/N6NB, published a description of an antenna he called the "Quagi."[13] This antenna is basically a Yagi with a cubical quad-type reflector and driven element. The design shown was low in cost, used a wooden boom, and was fed directly with coax. Newer designs have been published;[14] DL9KR and others have further optimized this design and modified the feed systems for use in large (groups of 16) EME arrays. This design could probably still use some optimization; only a few specific designs are available.

the log-periodic antenna array

The log-periodic antenna array,[15] a series of elements resembling a Yagi antenna, but with all elements fed by a special feed system, has never gained much acceptance among VHF/UHF'ers (except ATV'ers) because it is basically a wideband (multi-octave) structure with only moderate gain. The late Oliver Swan developed a hybrid antenna called the log-periodic Yagi, which has somewhat wider bandwidth than the typical Yagi.[16] It uses a log-periodic feed system to excite the directors of a Yagi structure. A 70 cm log-periodic Yagi has also been published.[17]

loop Yagis

Until recently, very few Yagi designs were used on the 23 cm band because they were hard to duplicate and extremely tight tolerances were required. The boom-to-element attachment had been a problem both electrically and mechanically, which further aggravated the tolerance problem.

In 1974, Mike Walters, G3JVL, decided to take a different approach to the Yagi antenna. At first he experimented with cubical quad loops made of wire, but could not obtain the desired performance. He then changed the elements to thin 0.028 inch (0.7 mm) aluminum straps 3/16 inch (4.7 mm) wide bent in a circle. After much trial and error, he developed a 23 cm, 28 element high gain "loop Yagi."[18] Later he developed longer models and also a table of corrections for different boom and element sizes.[19] The use of a round boom with his unique element-to-boom mounting method and the correction tables has been one of the reasons this design has been so successful. This antenna has been used as low as 70 cm and as high as 3 cm (10.256 GHz). A 45-element W1JR-designed loop Yagi had greater than 19 dB gain over a dipole as measured at the 1983 Eastern VHF/UHF Conference.

stacking

When really high gain is desired (such as in EME), Yagi antennas can be either arrayed or stacked. Generally speaking, every time you double the number of Yagi antennas in an array, you can increase the gain approximately 2.5 dB *if the proper stacking distance is used and the feedline loss is kept low* (this will be covered in a later column). Recent work on 70 cm EME tends to confirm that the best compromise is to use the highest gain Yagi design possible. Then the number of Yagis in the array will be at a minimum, and feedline problems will be more manageable. For general terrestrial operation, it is usually desirable to stack antennas vertically so that the beamwidth in the azimuth plane will stay wide and thereby spare you severe rotator aiming problems. This is particularly desirable for

meteor scatter communication, in which the field of reflection can be wide.

parabolic dish antennas

The parabolic dish (**fig. 3**) is still one of the favorite antennas on 23 cm and above — especially on EME. The gain of this type of antenna is primarily determined by its size (diameter), providing that the feed or illuminator is properly chosen. There are many advantages to using this type of antenna. It is frequency-independent, meaning that you can change frequency by simply replacing the feed

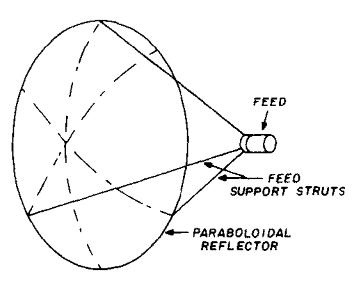

fig. 3. Typical parabolic dish (TNX W2IMU).

system with one for the proper frequency. A properly built parabolic dish antenna system will also be quieter on receiving because side and rear lobes are lower than most other antenna systems. Its principal disadvantages are that efficiency is low (typically 50 to 55 percent, even if everything is done right) and high wind resistance. (Recommended feed systems for parabolic dishes are described further in reference 3.)

conclusion

Hopefully the above information and the additional articles referenced will make it easier for you to determine which antenna type is best for you. The collinears are usually low in cost and easy to build, but if high gain is desired, they must be large. Where multiple band capability is desired on a single mast, the Yagi is hard to beat. When high gain on UHF and especially EME are sought, the parabolic dish offers many advantages; but again, if high gain is desired, the antenna system must be large. While we've discussed many types of antennas for VHF/UHF, we haven't discussed them in great detail. We'll deal with each type more comprehensively as time passes. But it will be up to you to weigh the advantages and disadvantages of each, make your decision, and take the plunge.

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VHF/UHF coming events:
Best EME Weekend: February 18-19

ham radio

the weekender

transmitter tuning aid: buffer your load with this resistive network

The **temporary high SWR** often encountered when tuning up your transmitter and tuning unit can often lead to disastrous consequences: abnormally high voltages and/or currents can easily damage expensive equipment, even in only a few seconds. If you once, perhaps, pit the variable capacitors of your final amplifier by an arc over, there is very little that can easily be done to repair the damage. In some cases, manufacturers have designed automatic circuitry that reduces power output when the SWR goes up, but this technique offers only limited protection rather than a basic solution.

With these concerns in mind, I developed and built a dummy load to overcome these problems. All discussions and calculations are based upon a 50-ohm impedance system, which is common to most of our modern transmitters and tuning systems. However, the equations can be used for any impedance system, or for any SWR limiting value desired.

The basic idea is to use a balanced-T resistive network, with a characteristic impedance of 50 ohms, to absorb most of the transmitter power during tune up, while still providing sufficient power to allow for proper tune up of your transmitter and antenna system. The isolation effect of the network is such that it limits your transmitter SWR to a maximum of 2:1, even if you were to short out or open the input to your tuning unit, which would be the worst condition that could develop.

Fig. 1 shows the basic circuit, along with theoretical calculated values of resistance, and wattage dissipation for a transmitter with an output of 100 watts. This figure was chosen for ease of calculation and presentation of the developed equations. Although the network is a balanced type, so that R_A is numerically equal to R_D in ohms, the different subscripts are shown to denote different wattage dissipation.

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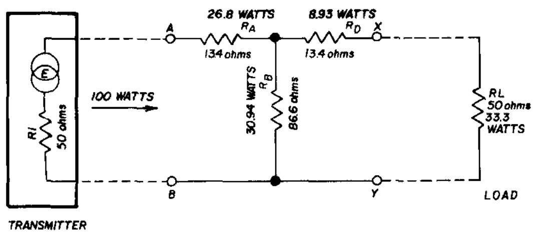


fig. 1. Balanced-T power-resistive network calculated values.

The equations relating the maximum value of SWR, which was chosen as 2:1, the characteristic impedance of the system R_C of 50 ohms, and the network resistance values are:

$$R_A + R_B = (R_C)(SWR) = (50)(2) = 100\ ohms$$

$$R_A + (R_D)(R_B)/(R_D + R_B) = R_C/SWR$$
$$= 50/2 = 25\ ohms$$

Having two equations and two unknowns, the network is solved using the basic quadratic equation, where:

$$R_A = R_D = 13.4\ ohms \text{ and } R_B = 86.6\ ohms$$

Again looking at **fig. 1**, if terminals X-Y are shorted out, the resistance looking in at points A-B is 25 ohms, so the transmitter SWR would be 50/25 = 2:1. If terminals X-Y are left open, the resistance looking into points A-B would be 100 ohms, and the transmitter SWR will once again be 100/50 = 2:1.

The characteristic impedance of the network is determined by these open and short circuit conditions where:

$$R_C = \sqrt{(R_{s.c.})(R_{o.c.})} = \sqrt{(100)(25)} = 50\ ohms$$

All of the original design requirements have been satisfied.

During actual tune up, R_L, the 50-ohm load resistance, is replaced by the input of the matching unit in the circuit in **fig. 2**.

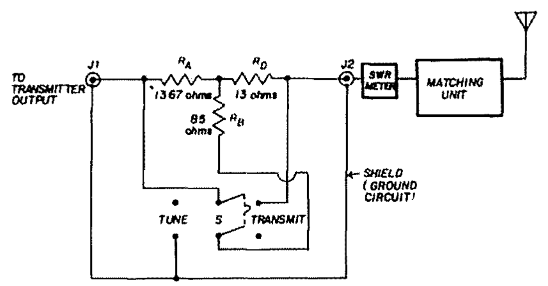
From **fig. 1**, it is seen that the power delivered to the load is 33-1/3 watts, or 1/3 of the transmitter output power of 100 watts at terminals A and B. This still provides plenty of power for tune up, yet offers an added advantage in that your "tune up QRM" has dropped by a factor of $(10)(log_{10})(1/3) = -4.77\ dB$. This reduction will be appreciated by other Amateurs on our already crowded bands.

A few years ago I experimented with building and using ordinary two-watt carbon resistors in parallel,

immersed in oil for a simple 50-ohm dummy load. I decided to follow the same procedure used then for the final T network, (**fig. 2**), in this project. My earlier tests had shown it was possible to use a heat dissipation factor of 2.5 as a conservative approach for normal tune-up time. To choose resistors, I checked to see what was available at my local electronics store, and then calculated what was needed to approach the theoretical values previously identified. For R_A, six 82-ohm, 2-watt carbon composition resistors in parallel equal 13.67 ohms with an allowable wattage rating of *(watts per resistor)(number of resistors) (derating factor) = (2)(6)(2.5) = 30 watts*, which was above the previously calculated value of 26.8 watts. (The values of resistors specified are shown in the parts list of **fig. 2.**)

Although the actual network resistance values differ slightly from the theoretically calculated values, actual tests showed the differences to be insignificant in operation. The network is very tolerant of small resistance deviations. This means you can simply use off-the-shelf items — there's no need to handpick resistance values.

There are other advantages in using the unit. Because your transmitter SWR during tune up can never go above 2:1 under the worst of conditions (and even then your reflected power is still only 11 percent of forward power), *your final amplifier output tuning control positions will be very close to their actual positions when you are finally tuned up to a perfect SWR of 1:1*. This makes tune up faster than it would be if you tried to tune up your amplifier while it was connected directly to your matching unit. Under the latter circumstances, the operation of tuner and transmitter controls is often so complex as to make one wish for four hands instead of two, and an extra pair of eyes to watch the SWR meter, lest it go much over 2:1.

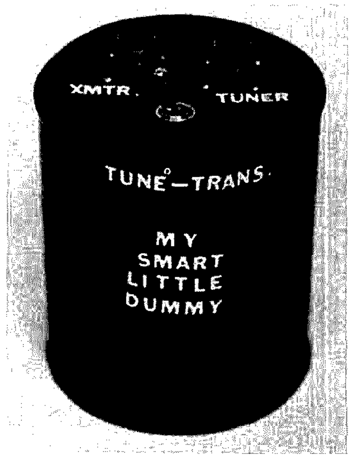


J1,J2	SO-239 coax connectors
R_A	82-ohm, 2 watt, carbon resistors — 5 percent tolerance (6 needed)
R_B	680-ohm, 2-watt, carbon resistors — 5 percent tolerance (8 needed)
R_D	39-ohm, 2-watt, carbon resistors — 5 percent tolerance (3 needed)
S	10 amp, 250 volt, DPDT toggle switch

1 quart paint can (empty)
1 quart transformer oil, mineral oil, or automobile transmission oil

Caution: Transformer oil of unknown origin may contain PCBs, a hazardous compound. See "PCBs in the ham shack," ham radio, December, 1983, page 42.

fig. 2. Completed balanced-T circuit values and placement in output line.



Completed balanced-T resistive network.

Another advantage of the unit is that you can switch it from TUNE to TRANSMIT when you are tuned up without having to turn off transmitter power, the circuit design prevents dangerous switching transients from occurring. Finally, because the unit is an in-line device, you will not need an expensive coax switch as might be necessary with a regular dummy load.

Although the design shown was for a balanced-T network, an equivalent Pi network with suitably calculated resistances and wattage values works just as well. The technique is also adaptable for high power. Actually, my first unit built was a converted home-brewed, flea market, 1 kW gallon-sized dummy which used a number of vitreous enameled non-inductive resistances, similar to the ones manufactured by Dale Electronics, Inc., P.O. Box 609, Columbus, Nebraska 68601. Although my copy of their catalog did not specify the highest frequency at which their non-inductive resistors could be used, my old re-structured dummy, with similar resistances, worked fine down through 10 meters.

This device has also been thoroughly tested by Russ Forsyth, K4YS, under a variety of operating conditions. His helpful suggestions and comments are part of this article.

ham radio

Correct carrier oscillator adjustment,
good linearity
and the proper microphone
make the difference

better-sounding SSB

If you want your SSB transmit audio to sound like your voice, this article was written for you.

I have never accepted the notion that good audio died with AM. After all, isn't SSB just AM with two of the unnecessary parts removed? The question is, why doesn't SSB audio give you good audio quality all the time? Armed with some electronic test equipment and a yearning for nice sounding audio, I set out to find the answers.

The results of my research indicate that three criteria must be met to achieve transmit audio quality that compares favorably with AM: the first is smooth frequency response; the second, a wide enough passband or "window" to be able to include most of the important frequencies in a human voice; and third, distortion should be low enough so that the voice does not sound rough.

generating SSB

In most SSB transceivers the mixing of the voice (audio) frequencies and the RF takes place in a balanced modulator. The balanced modulator is really a mixer that combines the audio and RF together. However, mixing is not an accurate way to describe the process, since the output is the sum of the audio frequency and the RF, or carrier frequency, and the difference of the audio frequency and the carrier frequency. (The name "mixer" should probably have been "adder-subtracter.") The *sum* output is upper-sideband (USB). The *difference* output is lower-sideband (LSB). Since both USB and LSB come out together on the sole output terminal, the result is double-sideband (DSB). Because of the balanced design of the modulator, neither the audio nor the carrier oscillator (input) signals should appear in the output.

It is important to remember that USB is always an addition process, and that LSB is always a subtraction process. For example, with a carrier frequency of 10,000,000 Hz and an audio frequency of 440 Hz, the USB output is 10,000,440 Hz and the LSB output is the difference or 9,999,560 Hz. If a second audio frequency were also fed into the balanced modulator along with the 440 Hz signal, then two frequencies would appear on USB and two frequencies would appear on LSB. If 1000 Hz were included with the 440 Hz signal, the USB would contain two frequencies: 10,000,440, Hz and 10,001,000 Hz. LSB would also contain two frequencies: 9,999,560 Hz and 9,999,000 Hz. When a voice is applied to the input of the balanced modulator, the output becomes more complicated, since a human voice contains many frequencies, but the additive and subtractive processes still apply.

producing just one sideband

The most popular system of selecting only one sideband incorporates a filter to reject the unwanted sideband and pass the desired sideband. It is common practice to use only one filter for both sidebands. In order to do this, two different carrier oscil-

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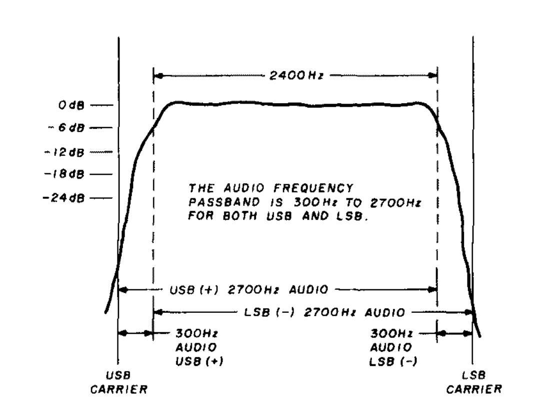

fig. 1. A typical 2400 Hz SSB filter with carrier oscillator frequencies set 300 Hz from filter window edges.

lator frequencies are used. On USB the carrier oscillator is placed about 300 Hz lower than the low frequency edge of the filter window. Since USB adds the voice and carrier frequencies, the sum will be higher than the carrier frequency and any audio frequency of 300 Hz or more will create USBs that can pass through the filter window. On lower sideband, the carrier frequency is placed about 300 Hz above the upper edge of the filter. Since LSB subtracts the audio frequency from the carrier frequency, an audio frequency of 300 Hz would just reach the upper edge of the window. In either case an audio frequency of 200 Hz would be mostly rejected since it could not add or subtract enough to each carrier frequency to reach the window. If the carrier oscillators were reset to a point 200 Hz from the edges of the window, then 200 Hz would produce a signal that will pass through the window. It is important to remember that the audio frequency itself does not pass through the window. The signals that pass through the filter are at a radio frequency, usually around 9 MHz. These signals are offset from the carrier frequency by the audio frequency. In order to receive SSB, the signal is combined with the same carrier frequency in the receiver as was used originally in the transmitter. The combining is done in another mixer called a product detector which produces audio output. This is the reverse of the process that began in the balanced modulator.

what voice frequencies are needed?

In the above example with the carrier oscillator frequencies set 300 Hz from the upper and lower frequency edges of the filter, it was shown how a 300 Hz audio signal would just reach the filter window. The next point to consider is how high in frequency can the audio signal go without missing the filter window. This is an *important* consideration because human speech contains high frequency sounds that are made by using the tongue to direct airflow against the teeth. This group of sounds, known as sibilants, includes S, soft C, F, and H. To do real justice to human speech, the window should allow at least 2800 Hz to pass. Allowing only 2500 Hz to pass makes a noticeable difference in the clarity of the sibilants. If a standard 2400-Hz wide filter is used, then the opposite side of the window is 2400 Hz away from the edge of the filter where the frequency generated by the 300 Hz audio signal would pass (**fig. 1**). The highest frequency audio that could pass without falling out of the window would be 300 *plus* 2400 or 2700 Hz. This audio would be acceptable, but would have more sibilance if the carrier frequency were moved 100 Hz farther from the filter window edge. This would move the audio passband from 400 Hz to 2800 Hz. This audio would be pleasing to most people with normal hearing. If a wider filter window were used — 2700 Hz — the audio quality would sound even better, since the carrier oscillator frequency could be reset so that the full range of important human speech frequencies — 300 Hz to 3000 Hz — could pass through the 2700 Hz window.

linearity

If the audio sounds rough and splatters on other frequencies, then linear operation is not occurring. The human ear can detect low levels of distortion. Some musicians can hear less than one part distortion in 10,000. That is −40 dB. I don't notice on-the-air distortion if it is −38 dB or better. If a station 4 kHz down the band has a 30 dB over S9 signal and his distortion products are down only 30 dB, you are going to see close to S9 splatter on the window you are listening to.

Most of the radios on the market today are fairly clean. The cleanest by far are those with vacuum tube finals employing RF negative feedback. If operated conservatively, some of the solid-state radios can also deliver a clean sound.

On my solid-state transceiver I was able to achieve −40 dB distortion products on my voice, but only after carefully setting the driver and final transistor idling current to the values called for in the service manual. I could only maintain the −40 dB distortion level if I kept the ALC level in the bottom quarter of the ALC scale. If I tried to operate with the ALC at the top of the "safe" range, the distortion increased by a factor of nine.

This test does not apply to the common two-tone method of measuring distortion. A human voice is much more complex than two tones and provides a

more accurate and severe performance test. Blowing into three holes of a harmonica also works well.

The mechanism of non-linearity is essential if a mixer is to do its job of adding and subtracting. In an amplifier, non-linearity means that the amplifier is going to do some unwanted adding and subtracting of any two or more frequencies fed into the amplifier input. Human voices have many frequencies. If the linear amplifier is not linear, there will be more frequencies, and some of them are likely to land on someone else's conversation on a nearby frequency.

measuring distortion

Distortion can be measured with a separate receiver, provided it has an S-meter of known accuracy. The linearity of the S-meter is the essential factor in making this test. The number of microvolts that equals S9 does not matter. Use a short clip-lead for the receiver antenna and a 20 dB attenuator. Tune to the frequency on which you are transmitting and watch the S-meter while you speak into the microphone. Note the S-meter reading and retune the receiver either 4 kHz above or 4 kHz below the transmitter frequency. Speak again and note the S-meter reading. The dB difference is your distortion level. A reading of −40 dB is excellent. A reading of −30 dB is not quite anti-social, but −20 dB is definitely in the "skunk" category. The all-time record worst distortion I have witnessed was −9 dB — this is *not* the way to win friends. Measuring distortion as described works best on your own transmitter. It is possible to use the same method on the air to measure someone else's distortion, but fading signals can make measurements inaccurate. To do this, a real antenna is more useful than the clip lead. I recommend 4 kHz spacing to ensure that your receive window does not appreciably overlap the other operator's transmit window. If the windows touch, the measurement will mean nothing, since you won't be able to separate the distortion from the desirable audio frequencies.

design defects

During my first SSB experiments I was amazed to learn that listeners could detect a small change in flatness of frequency response. If your transmitter has a 3 dB loss at the high end, the lack of sibilance is noticeable. Many of the modern rigs on the market today have a built-in rolloff of the high frequency transmit and receive audio. The amount of high frequency (treble) rolloff varies from 4 dB to 10 dB in the rigs that I have measured. Usually the receivers have the more uneven audio response.

The usual culprit in the transmitter section can be found at the collector of the microphone amplifier. In my transceiver there was a 0.015 microfarad capacitor from the collector to ground. About 60 percent of my 2900 Hz audio was going to ground at this point. I replaced this with a 0.001 microfarad capacitor and the treble passed through unscathed. Since the purpose of this capacitor is to prevent your own RF from getting into your audio, a non-inductive capacitor such as a disc ceramic type is advised.

I noticed that the high frequency audio in my receiver seemed attenuated. My test equipment confirmed a 9.5 dB rolloff at 2900 Hz. This meant that 11.2 percent of the audio was making it to the speaker at that frequency and that 88.8 percent was not. The culprit in this case was a large-value capacitor at the output of the product detector. The value used was 0.033 microfarad. This capacitor had a reactance of only 1600 ohms at 2900 Hz. The audio impedance of the product detector was about 5000 ohms. Shunting a 1660 ohm capacitor to ground is bound to be a spoiler. The purpose of this capacitor is to let the audio pass and to suppress the RF energy. Again the capacitor is necessary — but the value must be picked with some prudence. (Bigger isn't always better.) I changed the capacitor to 0.0033 microfarad. The audio sparkled. Voices were easier to understand.

microphones

Anyone who reads the advertising for microphones knows that you need a special microphone for SSB. It sounded reasonable to me, so I bought two microphones with lots of "punch." One was a crystal element type. The other microphone had a rising response dynamic element. Both microphones had more highs than lows — just the ticket to "get more DX." I also owned a flat response electret condenser microphone.

Early one winter morning I was working a DX station on 80 meters. I asked the DX station if he would listen to three different microphones and select the one that was easiest to copy under adverse conditions. I tried microphone number 1, number 2, and number 3. The microphone that produced the best copy was microphone number 3 — the condenser microphone — and the same microphone my local friends preferred to listen to. To be sure, this was only one test, but it certainly cast doubt on the traditional view of what it takes to be understood on SSB. Perhaps we don't really need to sound like ducks after all. Many amateur radio transceiver manufacturers now offer condenser microphones. Most of the microphones worn by television newscasters are of this type.

adjusting the carrier oscillators

The most important adjustment in an SSB transceiver is the carrier oscillator. This adjustment sets

the position of the audio frequency window on both RECEIVE and TRANSMIT. The only test equipment needed is a common audio generator that can be adjusted for a few millivolts and an RF wattmeter-dummy load. Connect the generator to the microphone input. Adjust the audio frequency for 1000 Hz. Set the level at a value that is about half of the maximum output power — for example, 40 watts. Adjust the generator higher in frequency until the wattmeter reads one quarter of the wattage produced when the audio frequency was 1000 Hz. This 25 percent power point is the minus 6 dB rolloff. Move the audio frequency lower until the power drops to 25 percent again. All of the frequencies between these two audio frequencies comprise your transmit window. If you find that your high frequency term is less than 2800 Hz at the rolloff point, your carrier needs to be moved farther away in frequency from the filter window. The low frequency rolloff should be 300 Hz or higher, but not above 400 Hz or it will sound thin (or tinny). If you are using a 2000-Hz wide filter you have a problem unless you want to sound like a 1910 model telephone. Wider filters are available for many rigs, direct from the manufacturer. I bought 2700 Hz filters for both of my transceivers.

Some people believe that a wider filter will impair reception because of increased splatter from adjacent stations. Most of the splatter on the ham bands is due to distortion products coming from the other fellow's transmitter; distortion products such as these can never be eliminated with a receive filter because the distortion is on the same frequency as the station you are trying to listen to.

There are devices on the market that will correct for a slight loss in treble transmit response caused by a 15-cent capacitor being the wrong value. But if the carrier oscillator is misadjusted, these devices will not be effective. For example, if the carrier oscillator is set so that the high frequency rolloff is 2200 Hz, the SSB filter is going to have the last word on what gets through. Boosting the treble with an accessory can make the problem less noticeable, but will not restore the high frequency rolloff to 2800 Hz as it should be.

Carrier oscillators can also be adjusted by ear. The rule is to simply set it to where it sounds good. While this may take a little longer than it would using test equipment, the method works. The only problem lies in finding a listener who is not afraid to tell you what you really sound like so you can know which way to move the adjustment. Some people don't want to hurt your feelings; they don't like to give a "bad" report.

Clean audio is a pleasure to listen to on the air, and rotten, splattering audio is a blight on the bands.

ham radio

ham radio TECHNIQUES

Bill Orr W6SAI

the groundplane loop antenna

Every once in a while an interesting antenna comes along that causes me to ask, "Why didn't I think of that idea?" A case in point is the groundplane loop antenna described by Hans Wuertz, DL2FA, in *cq-DL*, the monthly publication of the German Amateur Radio Club. An English translation of the pertinent data on the loop was recently published in *Radio Communication*, the journal of the Radio Society of Great Britain (RSGB).

The basic configuration of the DL2FA loop is shown in **fig. 1**. The design is a half of a full-wave loop antenna with the other half formed by the ground image. With a 0.2 wavelength half-loop for the lowest operating frequency, DL2FA claims that the antenna can be tuned to cover a 2-to-1 frequency range (3.5 to 7 MHz, for example).

Even though the radiation resistance of the antenna is low, the efficiency of the antenna is excellent, and at 3.5 MHz the field strength is only 2 dB less than a full-size quarter-wave vertical antenna. At 7 MHz, the field strength is within 0.5 dB of a full-size vertical.

To achieve low loss, the loop and ground return are constructed of 3/8-inch (approximately 10 mm) diameter copper tubing. The copper

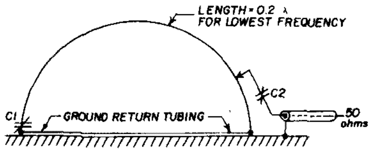

fig. 1. The DL2FA groundplane loop with gamma match feed. Tap point is approximately 0.04 wavelength. Ground return tubing is soldered to ground screen. Semicircle length is 27 feet 6 inches for 80-40 meter operation.

tubing that runs along the surface of the ground is soldered to a groundplane or mat made of chicken-wire (size not specified).

Measured data for the groundplane loop is tabulated in **table 1**.

FIBS revisited

A few years ago 80-meter operators in southern California and Arizona were surprised to hear one of the rarest broadcasting stations in the world: the *Falkland Islands Broadcasting Station* operating on 3.958 MHz. The signals would appear on an otherwise dead band in the late afternoon hours for a few moments and then disappear. The signals seemed to peak in the spring and fall.

Reports indicate the station is being heard once again. It seems that they are using single sideband with carrier, with the audio present on the upper sideband. Reception reports on this unusual station would be appreciated. The programs consist mostly of pop and disco music with BBC news on occasion.

It is thought that the FIBS propagation into the Northern Hemisphere is by other than normal ionospheric reflection. One theory is that the signal (and other DX signals near that frequency) may be propagated by a "whistler" path which extends into outer space far beyond the ionosphere.

a simple 50 to 75 ohm RF transformer

Shown in **fig. 2** is a simple high-frequency transformer that provides wideband transformation between 50 and 75 ohm unbalanced lines. The design is by JA1AKP, and appeared in a recent issue of *CQ-ham radio* (Japan). The transformer is wound on a FT-114 (μ = 125) No. 61 ferrite core (Amidon FT-114-61) having an outside diameter of 1.142 inches and an inside diameter of 0.75 inch. A trifilar winding of No. 18 wire is used, with two coils having 8 turns and one coil having 4 turns.

A 100 pF mica capacitor to ground is placed at the junction of the two 8 turn coils and a 10 pF capacitor to ground is placed at the 75 ohm port. The units are 500 volt rated. The balun is designed for 3.5 to 30 MHz at the 100 watt average power level.

ancient modulation

Since the late 1950's single sideband has reigned supreme on the Amateur bands. Users of amplitude modulation ("ancient modulation") have been greeted with derision and have gradually retreated to obscure regions of 160 and 10 meters. Many newly-licensed Amateurs have never heard an AM ham signal!

While Amateurs may pride themselves that sideband has provided them with a quantum leap ahead in communications reliability, they may forget that amplitude modulation is still the prime communication medium in the world. After all, the broadcast band is full of AM signals, and there are a lot of them!

Generally speaking, while the quality of Amateur SSB voice transmission has deteriorated,* the quality of AM transmission has improved. The modern term for broadcast quality is "transparency" — the transmitting system must be "transparent" with regards to the modulating signal.

While much thought and ingenuity have gone into AM transmission in the last decade, most of it has bypassed the Radio Amateur. New techniques, such as high-level pulse-duration modulation, AM stereo and audio processing are unknown in the ham bands, but noise reduction systems, equalizers, delay/reverb effects, level controllers, low frequency extenders and other processing techniques are routinely used by AM broadcasters to make their audio more realistic and outstanding.

On the receiving end of the line, even the less expensive clock-radio combo receivers have audio systems that make the average ham receiver blush with shame.

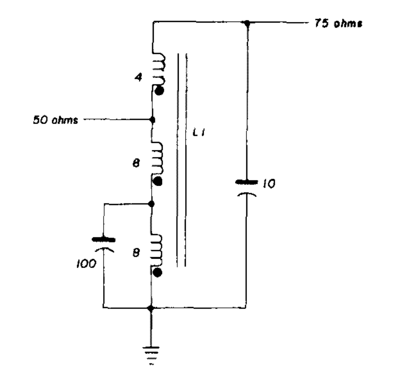

fig. 2. Wideband transformer to match 50 ohms to 75 ohms over the range of 3.5 to 30 MHz. Amidon FT-114-61 (or equivalent) core is used. Trifilar winding of No. 18 wire consists of two coils of 8 turns and one coil of 4 turns. See text for further details.

Even though most ham signals are "strained" through a narrow-band IF filter, there is no reason why the rest of the receiver should be below par as far as audio quality goes.

Recently I put a little audio feedback around the output stages of my trusty ham receiver, replaced the 19-cent (?) output transformer with one that had some iron in it, and ran the whole works into a spare stereo speaker. My goodness! I didn't recognize the sound of the ham bands! A lot of signals (whose operators previously sounded as if they were talking into a tin can at the end of a long string) actually sounded quite pleasant, and some signals sounded downright natural!

I then wondered what my transmitted audio quality was like. I put a good audio generator into the transmitter and looked at the audio signal as it entered the balanced modulator. I didn't like what I saw. Any resemblance to a sine wave was minimal.

It was no job at all to add a feedback resistor and to change another resistor in the low-level audio stages to make the waveform look very passable. Now, what came out of the exciter audio system bore a close similarity to what went in.

At this stage I stopped as I became bogged down into an interminable discussion of microphones with friends who knew a lot more about them than I did. The consensus was that the best way to improve audio quality was to throw away the microphone that came with the exciter and buy a good one! Without exception, all of my friends who had particularly pleasing and "punchy" signals on the air had replaced their standard transmitter microphones with different (and more expensive) units.

So there it is. The "ancient modulation" techniques of 1984 and the techniques used by up-to-date broadcasters certainly have some merit for use in the Amateur bands. Which of these new techniques should we examine? I throw the column open to discussion from the readers. (What's the input from hams who work in broadcasting? I'd especially like to hear from *you.*)

table 1. Groundplane loop dimensions and performance data. (Courtesy of *Radio Communication*, published by the Radio Society of Great Britain and available from Ham Radio's Bookstore, Greenville, New Hampshire 03048.)

band	semicircle length	C1 (pF)	bandwidth (−3 dB points)	efficiency % (MHz)	comparison with 1/4 wave vertical
3.5 - 7.0	27 ft 6 in (8.4 m)	332	5.7 kHz	70 (3.5)	−1.91 dB
			to 67 kHz	96 (7.0)	−0.55 dB
7.0 - 14.0	13 ft 9 in (4.2 m)	184	11 kHz	77 (7.0)	−1.52 dB
			to 147 kHz	97 (14.0)	−0.50 dB
14 - 28	6 ft 10 in (2.1 m)	102	24 kHz	83 (14.0)	−1.22 dB
			to 323 kHz	98 (28.0)	−0.47 dB

staggered stacking

It is common practice in the VHF bands to stack two or more high-gain Yagi beams to obtain increased gain. Adjustment of the stacking distance can improve sidelobe reduction of the array.

Experiments conducted by G.J. McDonald, VK2ZAB, and reported in the Australian magazine *Amateur Radio*, indicate that an improved

*See "better-sounding SSB," page 58 — Editor

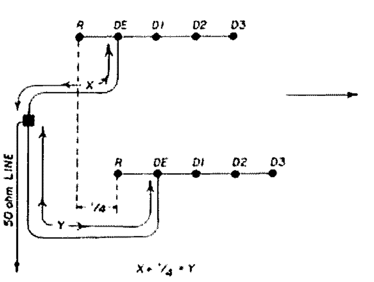

fig. 3. The VK2ZAB array employing staggered stacking. Bottom Yagi is advanced one-quarter wavelength in front of upper array and extra quarter wavelength is added to the phasing line. Front-to-back ratio is improved by this modification.

front-to-back ratio in a Yagi stack can be achieved by staggering the antennas. That is, the bottom Yagi of each vertically stacked pair is advanced one-quarter wavelength on its mounting so that it projects one-quarter wavelength in front of the upper array (fig. 3). The phase lead in the forward direction is cancelled by adding an extra quarter wavelength to the phasing harness connected at the bottom Yagi, thus insuring that the current to both arrays arrives from the feed junction in phase.

Signals arriving at the rear of the array are subject to a 90-degree lag by virtue of the position of the bottom Yagi relative to the top one, plus the additional 90-degree lag caused by the extra length of phasing line.

Results? With a conventional stack of two Yagis, the measured front-to-back ratio was about 12.5 dB. The staggered stack provided a front-to-back ratio of 24 dB, for an improvement of 11.5 dB. And that is equivalent to about 2 "S" units — enough to make the difference between just being able to copy a DX signal through QRM from the rear of the array and copying it comfortably!

Final measurements showed the staggered stack had about 0.25 dB more gain than the conventional stack and also reduced the sidelobes to some extent.

ham radio

VHF meteor scatter communications

Sporadic meteor activity improves reliability

Meteors are space junk — typically, small bits of stone or metal from a passing comet, or perhaps leftovers from the formation of our solar system.

When a meteor enters earth's atmosphere it almost always burns up. Its violent passage produces heat, light, and an ionized trail of gases in the upper atmosphere which reflect or reradiate radio signals, an effect known by those who observed the return of American astronauts from space. Not long ago, the crews of the space shuttle *Columbia* lost contact with the ground for several minutes during reentry because of the reflective properties of ionized gases created by the tremendous heat of reentry.

A few Amateurs have taken advantage of these ionized trails to communicate beyond the VHF horizon, using various modes of transmission. Until now, however, communications using the reflective and refractive qualities of meteor trails have been restricted to periods of stream or shower meteors, because only these periods provide a level of ionization high enough to support two-way communication using Morse code or voice modes.

Once again the Amateur community stands on a technological frontier, for with the aid of today's microprocessor systems, some of the more than 100 million meteor trails that occur daily can be used even though their peak reflective trail lasts only several milliseconds.

Perhaps the most appealing aspect of the meteor-burst concept is the use of bands "closed" to normal communication because of poor conditions. Sporadic meteors occur 24 hours a day, 365 days a year. This means you need not store that 6-meter rig in the closet during minimal sunspot activity; you can go right on using it to communicate in a new way using new transmission techniques.

shower and sporadic meteors

Ionized paths created by the passage of sporadic and shower meteors allow two-way radio communication by returning a portion of the signal to the earth much the same as the ionosphere does at frequencies below 30 MHz. Current Amateur meteor-scatter operations take place mainly during large meteor showers. These predictable showers are composed of meteors large enough to cause trails lasting several minutes; the trails are long enough to allow operating techniques to take place at humanly controllable speeds.

Meteor showers or *stream meteors* occur at predictable times during the year (see **table 1** and **fig. 1**), with peak activity during June, July, August, and December. Communication using the shower meteors is a VHF speciality in the same class as using sporadic E, tropospheric scatter, and ducting conditions. (In contrast, the sporadic meteor-scatter user need not be concerned with meteor-shower schedules or optimum direction calculations.)

Because they appear to radiate from certain points

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in the sky, meteor showers are usually named for a constellation visible nearby: for example, the Geminids appear to come to Earth from the constellation Gemini. As the constellation moves across the sky, communication is established and maintained by keeping the antenna trained on Gemini as the Earth rotates.

Unlike shower-borne stream meteors, *sporadic meteors* bombard the earth's atmosphere constantly. An average of 100 million meteors large enough to leave an ionized trail enter the earth's atmosphere daily. The usable trails occur at an altitude of 50 to 75 miles (80 to 120 km) and must last more than a few hundred milliseconds (**table 2**) in order to be usable.

Sporadic meteors are the small meteors characterized by the "ping" noise heard by VHF DXers — hence the name, "ping jockey." Sporadic meteors are the ones used to communicate over distances up to about 1000 miles (1600 km) by relatively new data acquisition systems.[1]

Although sporadic meteors appear continuously, their most intense activity is between 0400 and 0800

table 1. Meteor shower data for 40 degrees north latitude. Peak days, meteor velocities, and hourly rates are given, in that order, under shower name.

shower and date (peak date) (velocity) (rate)	optimum time/path (local standard)	antenna offset direction	notes
January 1-4	0220-0740/NW-SE	SW	intense, but lasts
QUADRANTIDS	0740-0900/E-W	S	only several hours
Jan. 3, 42km/s	0900-1430/NE-SW	SE	
50/hr			
May 1-6	0410-0630/NE-SW	NW	
ETA AQUARIDS	0630-0830/E-W	N	
May 4, 64km/s	0830-1050/NW-SE	NE	
15/hr			
June 1-15	0515-0745/N-S	W	intense daylight
ARIETIDS	0745-0920/NE-SW	NW	shower, but mostly
June 7-8, 39km/s	0920-1015/E-W	N	small particles
60/hr	1015-1155/NW-SE	NE	
	1155-1425/N-S	E	
July 26-31	0100 /NE-SW	NW	
DELTA AQUARIDS	0100-0330/E-W	N	
July 29-30, 43km/s	0330 /NW-SE	NE	
20/hr			
August 10-14	2330-0430/NW-SE	SW	reliable; many
PERSEIDS	0430-0730/E-W	S	large particles
Aug. 12-13, 61km	0730-1230/NE-SW	SE	
50/hr			
October 18-23	0015-0145/N-S	W	
ORIONIDS	0145-0350/NE-SW	NW	
Oct. 21-22, 67km/s	0350-0500/E-W	N	
20/hr	0500-0710/NW-SE	NE	
	0710-0840/N-S	E	
November 14-18	0200-0430/N-S	W	Periodic, spectacular
LEONIDS	0430-0600/NE-SW	NW	every 33 yrs; poor in
Nov. 16-17, 72km/s	0600-0700/E-W	N	between. Next peak in
10/hr	0700-0840/NW-SE	NE	1999.
	0840-1100/N-S	E	
December 10-14	2115-0100/N-S	W	reliable, many large
GEMINIDS	0100-0200/NE-SW	NW	particles
Dec. 12-13, 35km/s	0200-0230/E-W	N	
60/hr	0230-0330/NW-SE	NE	
	0330-0715/N-S	E	
December 21-23	2200-1600/E-W	S	
URSIDS			
Dec. 22, 35km/s			
13/hr			

hours, local time, because this is the time when the local sector turns toward the direction of the Earth's travel through space. This is a solar system event, in that the occurrence is approximately the same at local time at any point on the Earth — just as daybreak or moonrise.

The peak in daily variation in sporadic meteor activity occurs when a sector of the Earth's surface becomes the "leading edge" of the planet's progress through space, effectively "leading" the sweep of the planet through space.

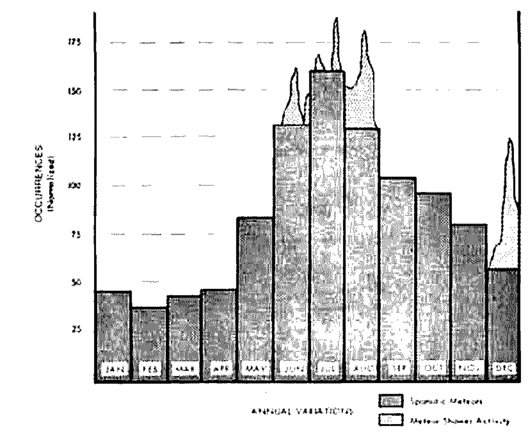

fig. 1. Normalized meteor activity by month.

From late afternoon through early evening, the "leading-edge" portion of Earth's surface changes to the "trailing edge;" only those meteors traveling fast enough to catch up are caught in Earth's gravitational field. Of these, only some will produce a usable trail. The late afternoon events are much further apart and meteors move much more slowly relative to the Earth's surface, resulting in greatly increased waiting time between usable paths (**fig. 2**).

Because the number of usable sporadic trails varies on an annual and daily basis, astronomers have concluded that sporadic-meteor materials are spread fairly evenly throughout the solar system, but confined to the plane that intersects the planets.

history

Mankind has been aware of meteors — so called "shooting stars" — since his first view of the night sky. The radio disturbance caused by meteors is a relatively recent discovery; Skellett,[2] in 1932, was the first to link an increase of noise in a receiver to the passage of meteors through the atmosphere. Allen[3] studied the meteor-scatter propagation of signals from an FM transmitter at Paxton, Massachusetts, in 1934. In this work, he identified both the shower and sporadic types of meteor trails. Virtually all advances in the field to date are based on his and Skellett's works.

The feasibility of meteor-burst communications was first tested by the National Bureau of Standards

table 2. Order of magnitude estimates of the properties of sporadic meteors.

	mass (grams)	radius		number of this mass or greater swept up by the earth each day	electron line density (electrons per meter of trail length)
particles pass through the atmosphere and fall to ground	10^4	3.15in	(8 cm)	10	10^{18}
Particles totally disintegrated in the upper atmosphere	10^3	1.57in	(4 cm)	10^2	10^{18}
	10^2	7.87×10^{-1}	(2 cm)	10^2	10^{18}
	10	3.15×10^{-1}	(0.8 cm)	10^4	10^{18}
	1	1.57×10^{-1}	(0.4 cm)	10^5	10^{17}
	10^{-1}	7.87×10^{-2}	(0.2 cm)	10^6	10^{16}
	10^{-2}	3.15×10^{-2}	(0.08 cm)	10^7	10^{15}
	10^{-3}	1.57×10^{-2}	(0.04 cm)	10^8	10^{14}
	10^{-4}	7.87×10^{-3}	(0.02 cm)	10^9	10^{13}
	10^{-5}	3.15×10^{-3}	(80 MICRONS)	10^{10}	10^{12}
Approximate limit of radar measurements	10^{-6}	1.57×10^{-3}	(40 MICRONS)	10^{11}	10^{11}
	10^{-7}	0.787×10^{-3}	(20 MICRONS)	10^{12}	10^{10}
	10^{-8}	0.315×10^{-3}	(8 MICRONS)	?	?

Note: Information from K. Davies, "Ionospheric Radio Propagation," National Bureau of Standards Monograph #80, April, 1965.

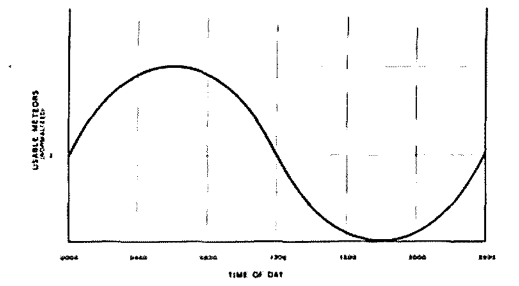

fig. 2. Normalized daily variations in meteor activity.

(NBS) between Cedar Rapids, Iowa, and Sterling, Virginia, a distance of 733 miles (1180 km).[4] A system built by NBS and operated from 1950 to 1952 showed the feasibility of long-range meteor-scatter communication in the 30- to 100-MHz band. Through experience with this system, NBS defined certain operational characteristics of meteor scatter.

Stanford Research Institute (SRI)[5] built the first true meteor-scatter communication system in 1956-57 between Bozeman, Montana, and Palo Alto, California. The SRI system used a two-frequency scheme with a burst rate of 454 bits per second (BPS). A pilot frequency containing no information was continuously transmitted. When the receiving station sensed the pilot signal, it would order the system to begin transfer of data. There were several problems with this approach; mechanical delays in the start-up of data transmission and low receiver sensitivity considerably limited the data transfer rate.

In 1956, the Canadian Defense Research Board[6] began using a system called JANET to conduct experiments over paths of 560 to 745 miles (900 to 1200 km). JANET showed a wide difference in hourly data transfer rates because of daily changes in the number of meteors per unit time and other ionospheric effects, such as sporadic E and aurora. JANET's overall average information rate was 25.5 BPS, which is equivalent to a speed of 38.25 words per minute (WPM).

The National Bureau of Standards resumed meteor communication experiments in 1959, combining JANET technology with some other innovations to attain an overall average data rate of 45 WPM with an error rate of 0.35 percent.[7] The same year, under contract with the United States Air Force, Hughes Aircraft Company developed a system for air-to-ground communication, introducing techniques for improved block transmission and error detection and correction.

In 1965, the North Atlantic Treaty Organization began operating a system called COMET (*CO*mmunication by *ME*teor *T*rails) over a 625-mile (1000-km) path using a 200 watt transmitter and a system for Automatic Repeat Request (ARQ).[8] ARQ is a system of error detection that allows the receiving station to

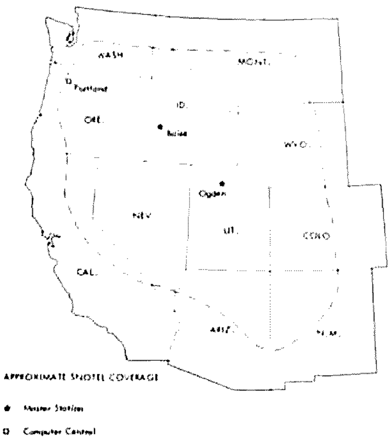

fig. 3. Approximate data collection coverage of SNOTEL system.

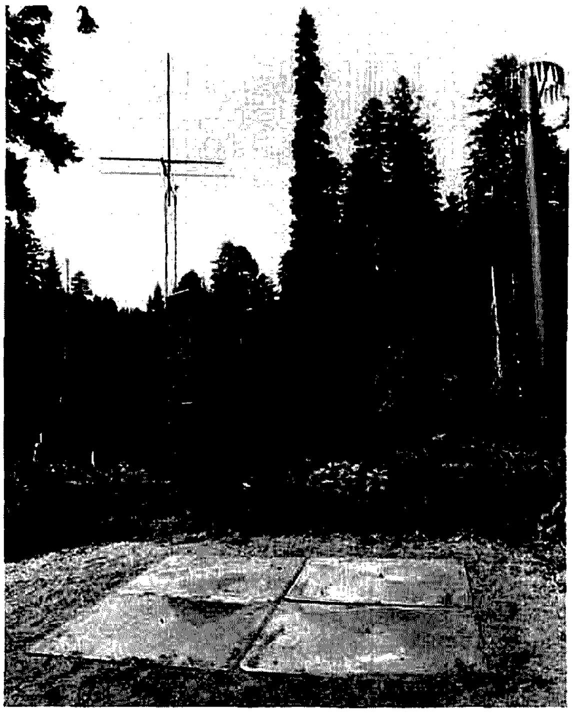

fig. 4. SNOTEL remote site with snow water sensors in foreground.

fig. 5. SNOTEL remote site in winter.

automatically request a repeat of the last information received. COMET achieves a worstcase data rate of 14 WPM and an average of 140 WPM.

largest meteor-scatter link

The Soil Conservation Service of the U.S. Department of Agriculture operates the world's largest meteor-scatter communications data collection system (**fig. 3**) installed over a period of four years (1976-1980), it is a hydrometeorological data collection system called SNOTEL (*SNO*w *TEL*emetry) located in the western United States.

Located in isolated mountain areas, the nearly 500 SNOTEL installations (**figs. 4** and **5**) collect and transmit detailed data on weather and snow conditions, updating information every 15 minutes. The data is used by the Soil Conservation Service for a number of purposes, including forecasting the amount of water available for irrigation, municipal water supplies, and power generation the following spring and summer.

SNOTEL is controlled by a computer in Portland, Oregon; the two master stations in Ogden and Boise (**fig. 6**) can request, receive, and acknowledge data from any of the remote sites, using phase shift keyed transmitters with Henry "4K ULTRA" finals and four phase locked loop-based receivers. Each of the four antennas at the master station is a five-element Yagi directed at a different quadrant of the sky and tipped up 30 degrees to optimize perfomance (**fig. 7**). Currently, the system collects data from a high percentage (typically 90 to 95 percent) of the more than 1015 active channels in the system during a 2-hour nominal polling period each day — a collection of 7000 characters or the equivalent of 1400 five-letter words per hour, and in maximum message lengths of 256 to 512 bits of sensor data.

possible Amateur applications

Current microprocessor technology opens up the field of meteor burst for the Amateur Radio community. An Amateur meteor-scatter system can be constructed using a microprocessor to control the receive and transmit activity. Literally *anything* that can be digitized — pictures, RTTY data — can be transmitted via meteor burst.

The key to an Amateur meteor-burst communication system is a standardized format, bit rate, and an "acknowledge" or ARQ system. Several versions of ARQ system introduced by current meteor-scatter technology are adaptable to Amateur Radio use. Microprocessor technology offers several schemes for developing ARQ systems for error detection and transmission requests.

fig. 6. Interior of master station at Boise, Idaho.

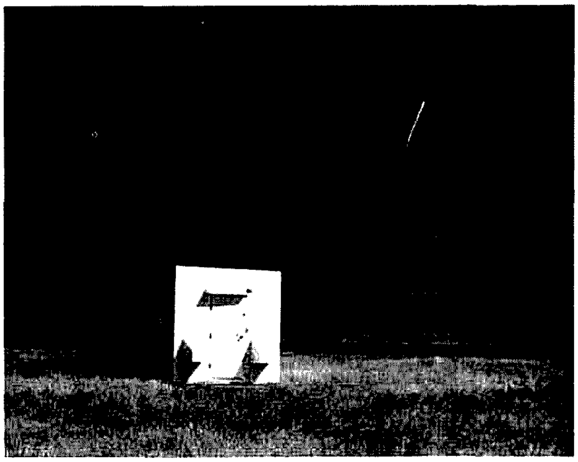

fig. 7. SNOTEL master station located at Boise, Idaho, and one of its four antennas.

A workable format, detailed in **fig. 8**, consists of 16-bit sync train (two 8-bit bytes), 72 words of data (characters), one byte of parity information, and one group number byte. (This is similar to the SNOTEL data stream and is convenient for computer use because of the 72 characters per line.) Inclusion of the group number allows the receiving end of the communications link to request repeat of a single burst or group of bursts. Basically this is a simple extension of the computer-monitored RTTY station that grew out of the Auto-Start-Auto-Answer RTTY station.

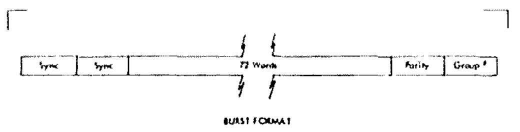

fig. 8. Possible burst format.

conclusion

Almost unlimited opportunity for discovery exists in the field of meteor burst communications. Areas open to Amateur experimentation include antenna design and direction control, error detection and correction techniques, and frequency dependence. Amateurs could take an active role in the improvement of remote control and RTTY systems and in perfecting techniques of data collection and monitoring.

With their access to technology appropriate for use in exploration of the field, Amateur Radio operators enjoy a unique opportunity for innovation and exploration in this expanding field.

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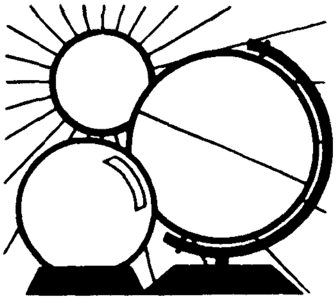

DX FORECASTER

Garth Stonehocker, K0RYW

last-minute forecast

Excellent DX conditions on the higher frequency bands (10-30 meters) are expected during the first and second weeks of the month due to a probable increased solar flux. A recurrent disturbed period is also possible around February 4, which should result in better transequatorial openings. Check with WWV's geophysical announcement at 18 minutes after each hour for confirmation of the high solar flux and geomagnetic A values.

Other periods of disturbance may be on the 14th and from the 22nd through the 28th. The lower bands should be very good all month and excellent during low solar flux periods in the third week of the month. Look for DX openings to unusual east-west locations during the disturbed days.

No significant meteor showers occur during February. The full moon and lunar perigee are on the 17th.

low band DX

Since publication of the nighttime bands in the propagation chart was begun, have you noticed how high those bands are, even at their diurnal minimum? DX stands for long distance contacts; even the shortest DX path, Europe, is 3800 miles (6200 km) from the East Coast. South America, Japan, and the Far East are approximately 6200 miles (10,000 km), while Antarctica and Australia are 9000 miles (14,600 km) and 9900 miles (16,000 km), respectively. The highest usable bands are related to the maximum usable frequencies (MUF) for that path. The longer DX paths are to the south across the equator where minimum MUFs are around 14-15 MHz now and will decrease by 3 MHz when the sunspot number drops to 16 (or the radio flux to 75 units). The other DX paths are in mid to high latitudes where MUFs are lower and paths are shorter, giving minimum MUFs of only 8-9 MHz. These will decrease by 2 MHz at the sunspot cycle minimum.

These numbers are based on monthly median (January) data. The daily solar flux factors from last month's *DX forecaster* can be used to forecast several days ahead. The *minimum* MUF is inversely related to the daily flux, decreasing with an increasing change in radio flux. Geomagnetic disturbances will also affect these minimum MUFs by 5 to 10 percent — the high latitude paths negatively and transequatorial paths positively.

Interestingly, even at sunspot minimum, 40 meters will be open to Europe, Japan, and the Far East and 30 meters on transequatorial paths. If you work 80 meters and 160 meters, they will be 50 percent and 75 percent below the MUF even on the short-northern paths. This doesn't mean they won't work; we know they do.

band-by-band summary

Ten and fifteen meters will be open for worldwide DX communication from sunrise until after sunset during the twenty-seven day solar flux maximums. Skip of 2500 miles (4000 km) (or multiples) is possible, and will follow the sun across the earth.

Twenty meters will be open to some area of the world for the entire twenty-four hour period on many days of the month. The band should peak in all directions just after local sunrise, and again toward the east and south during late evening hours. During hours of darkness the band will peak toward the west in an arc from southwest through northwest, encompassing Pacific areas.

Thirty meters is a daytime and nighttime band. The day portion should be like 20 meters except in that signal strengths may decrease during midday on days having high solar flux values. This band will also be useful well into the night and often through the night. Once again exceptions to this are on nights that follow very high solar flux value days. The problem time is usually the hour or so before dawn (diurnal MUF minimum). The workable distance may be expected to be greater than 80 DX at night and less than 20 during the day.

Forty and eighty meters will be the most usable nighttime DX bands. Most areas of the world will be workable from dusk until sunrise. Hops shorten on these bands to about 2000 miles for 40 meters and 1500 miles for 80 meters, but the number of hops can increase because signal absorption in the ionosphere's D-region is low during the night. The path follows the direction of darkness across the earth, similar to the way in which the higher bands follow the sun.

One-sixty meters will be similar to 80 meters, providing good working conditions for enthusiastic DXers who like to work nighttime and early morning hours, especially at local dawn.

ham radio

FEBRUARY

WESTERN USA

GMT	PST	N (ASIA FAR EAST)	NE (EUROPE)	E (S. AFRICA)	SE (S. AMERICA)	S (ANTARCTICA)	SW (NEW ZEALAND)	W (OCEANIA AUSTRALIA)	NW (JAPAN)
0000	4:00	20	20	15	10	15	10	10	20
0100	5:00	20	20	15	10	15	10	10	20
0200	6:00	20	20	15	10	15	10	10	20
0300	7:00	20	30	15	10	15	10	10	20
0400	8:00	20	30	15	10	15	10	10	20*
0500	9:00	20	30	20	10	15	10	10	20
0600	10:00	20	*30*	20	15	15	10	10	20
0700	11:00	*20*	*30*	20	15	15	10	15	20
0800	12:00	*20*	*30*	20	*20*	15	15	15	20
0900	1:00	*20*	*30*	*20*	*20*	*20*	15	15	*30*
1000	2:00	*30*	*40*	*20*	*20*	*20*	15	20	*30*
1100	3:00	*30*	*40*	*20*	*20*	*20*	*20*	20	*30*
1200	4:00	*30*	*40*	*20*	*20*	*20*	*20*	*20*	*30*
1300	5:00	*30*	*30*	*20*	*20*	*20*	*20*	*20*	*30*
1400	6:00	*30*	*30*	15	*20*	*20*	*20*	*20*	*40*
1500	7:00	40	20	15	15	*15*	*20*	*20*	*40*
1600	8:00	40	20	15	15	*15*	15	*20*	*30*
1700	9:00	40	20	15	15	15	15	15	*30*
1800	10:00	40	20	10	15	15	15	15	20
1900	11:00	40	20	10	15	15	15	15	20
2000	12:00	30	20	10	15	15	10	15	20
2100	1:00	30	20	10	10	15	10	10	20
2200	2:00	30	20	15	10	15	10	10	20
2300	3:00	30	20	15	10	15	10	10	20

MID USA

GMT	MST	N (ASIA FAR EAST)	NE (EUROPE)	E (S. AFRICA)	SE (S. AMERICA)	S (ANTARCTICA)	SW (NEW ZEALAND)	W (OCEANIA AUSTRALIA)	NW (JAPAN)	CST
0000	5:00	30	20	15	10	15	10	10	20	6:00
0100	6:00	30	20	15	10	15	10	10	20	7:00
0200	7:00	30	20	15	10	15	10	10	20	8:00
0300	8:00	30	30	15	10	15	10	10	20	9:00
0400	9:00	30	30	15	15	15	10	10	20	10:00
0500	10:00	30	30	20	15	15	10	10	20	11:00
0600	11:00	30	*30*	20	15	15	10	15	20	12:00
0700	12:00	*30*	*30*	20	15	15	15	15	20	1:00
0800	1:00	*30*	*30*	20	*20*	20	15	15	*30*	2:00
0900	2:00	*30*	*40*	*20*	*20*	*20*	15	20	*30*	3:00
1000	3:00	*30*	*40*	*20*	*20*	*20*	*20*	*20*	*30*	4:00
1100	4:00	*20*	*40*	*20*	*20*	*20*	*20*	*20*	*30*	5:00
1200	5:00	*20*	*30*	*20*	*20*	*20*	*20*	*20*	*30*	6:00
1300	6:00	*20*	*30*	*15*	*20*	*20*	*20*	*20*	*40*	7:00
1400	7:00	20	*30*	15	15	*20*	*20*	*20*	*40*	8:00
1500	8:00	20	20	15	15	*20*	*20*	*20*	*40*	9:00
1600	9:00	20	20	15	15	15	15	*20*	*40*	10:00
1700	10:00	30	20	10	15	15	15	15	*30*	11:00
1800	11:00	30	20	10	10	15	15	15	30	12:00
1900	12:00	30	20	10	10	15	15	15	20	1:00
2000	1:00	30	20	10	10	15	10	15	20	2:00
2100	2:00	30	20	10	10	15	10	10	20	3:00
2200	3:00	30	20	15	10	15	10	10	20	4:00
2300	4:00	40	20	15	10	15	10	10	20	5:00

EASTERN USA

GMT	EST	N (ASIA FAR EAST)	NE (EUROPE)	E (S. AFRICA)	SE (CARIBBEAN S. AMERICA)	S (ANTARCTICA)	SW (NEW ZEALAND)	W (OCEANIA AUSTRALIA)	NW (JAPAN)
0000	7:00	30	20	15*	15	15	10	10	20
0100	8:00	30	20	15	15	15	10	10	20
0200	9:00	30	20	15	15	15	10	10	20
0300	10:00	30	30	15	15	15	10	10	20
0400	11:00	30	30	15	20	15	10	10	20
0500	12:00	*30*	30	20	20	15	10	15	20
0600	1:00	*30*	30	20	20	15	15	15	20
0700	2:00	*30*	*30*	20	20	15	15	15	20
0800	3:00	*30*	*30*	20	20	20	15	*20*	*30*
0900	4:00	*30*	*40*	*20*	20	20	*20*	*20*	*30*
1000	5:00	*30*	*40*	*20*	*20*	*20*	*20*	*20*	*30*
1100	6:00	*20*	*40*	*20*	*20*	*20*	*20*	*20*	*30*
1200	7:00	*20*	*30*	*15*	*20*	*20*	*20*	*20*	*40*
1300	8:00	*20*	*30*	*15*	*15*	*20*	*20*	*20*	*40*
1400	9:00	20	*30*	15	*15*	*15*	*20*	*20*	*40*
1500	10:00	20	20	15*	15	15	*20*	*20*	*40*
1600	11:00	20	20	10	15	15	*20*	*20*	*40*
1700	12:00	20	20	10	15	15	15	15	*30*
1800	1:00	20	20	10	15*	15	15	15	30
1900	2:00	20	20	10	10	15	10	15	20
2000	3:00	30	20	10	10	15	10	10	20
2100	4:00	30	20	10	10	15	10	10	20
2200	5:00	30	20	15	15	15	10	10	20
2300	6:00	40	20	15	15	15	10	10	20

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during 'normal' hours.
*Look at next higher band for possible openings

EMI/RFI shielding: new techniques

part 2

Techniques and materials that reduce EMI

Part 1 of this two-part article served as a basic overview by identifying the problem of EMI/RFI, citing examples of its adverse effects, and providing a general discussion of EMI shielding effectiveness (SE). In one form or another, each of these various schemes involved plastics. This more detailed, more technical article examines methods of testing SE that employ RF techniques familiar to most hams.

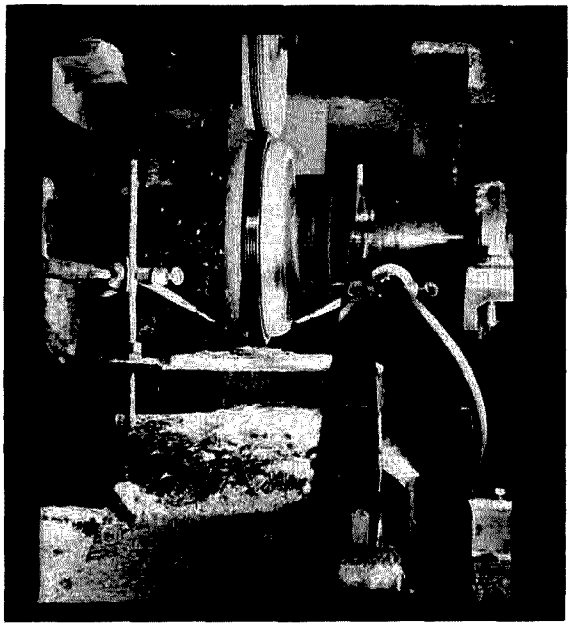

fig. 1. The generation of aluminum flakes.

Before examining SE and its testing, let's first examine the process of fabricating plastics in a manner that renders them effective for EMI shielding and containment, yet is economical enough to allow plastic enclosures to be used instead of metal ones.

EMI and plastics

The FCC tries to stringently enforce EMI emission standards that apply to electronic equipment. The pressures of foreign competition, however, demand that American manufacturers develop innovative cost-cutting manufacturing techniques such as using EMI-proof plastic enclosures instead of metal enclosures. But there are other ways than the use of enclosures to obtain or shield electronic devices from EMI/RFI; we can use metallic tape with adhesive backs (the adhesive is a plastic), neoprene and elastomer gaskets (both plastics), or sprays and paints in conjunction with plastics. In fact, plastics are so common that we sometimes overlook them, and often fail to fully appreciate or understand their versatility.

In order to better understand how plastics are used in potting and casting, in PC board conformal coating, in insulation, high frequency connectors and a myriad of other applications, let's begin by looking at materials used for shielding.

materials for shielding

Most EMI shielding processes in use today have disadvantages: precious metal powders are expensive; conductive carbon comes in but one color — black — zinc metallic sprays can lose adhesion, with small metal flakes falling onto PC boards and shorting out components; and paints are subject to abrasion and subsequent delamination when put through temperature extremes. What is the solution to the problem? The answer may be a new technique of utilizing structured fillers.

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Graphite fillers. With higher conductivity and better reinforcement characteristics than carbon blacks, graphite fillers look promising. Radar chaff, for example, originally developed to resonate at radar frequencies and to produce "phantom" targets for anti-aircraft batteries, has been successfully introduced into these types of composite plastics.

Impregnation of aluminum flakes into plastic is a relatively new process. A spinning wheel throws off flakes of molten aluminum alloy (see **fig. 1**); these flakes, with a tightly controlled aspect ratio (length-to-thickness ratio), are then deposited into a resin which causes them to become oriented in the direction of the flow of the resin. You have no doubt seen fingerlings or small minnows in schools orient themselves upstream in line with the flow of water; these flakes do the same in the resin. This is important because molding or extrusion processes require this. If all flakes are not oriented in the same direction, they experience lateral force and breakage occurs, with a subsequent reduction in their shielding effectiveness.

Typically, aluminum-impregnated plastics contain 18 to 22 percent flakes by volume because electrical conductivity depends not on the weight ratio of the flakes with their resin "holder," but on the volume ratio of flakes to resin. This means that a low-density polymer like polypropylene would need 46 percent weight, while a high-density polymer like PVC would require only 37 percent flakes to achieve the same 22 percent volume ratio.

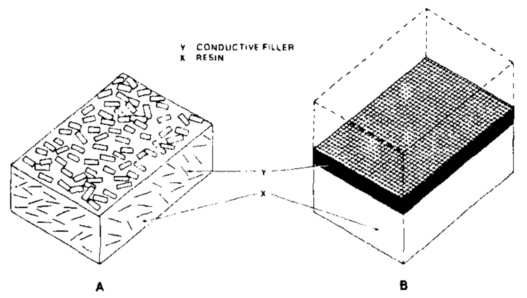


fig. 2. Composite of conductive particles suspended in an insulating matrix.

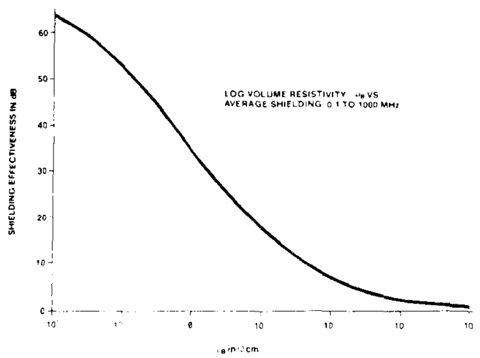


fig. 3. Graph of averaged surface and volume resistivity.

the acid test

To date, most testing techniques have been contrived to test conductors or pieces of pure conductive materials. But now, the purely resistive properties of a substance like straight wire no longer exist; the combining of metals with plastics has caused networks to result with both inductive and capacitive reactances.

In **fig. 2**, note that if probes were to pass through the insert on the left at precisely the right spot, they could miss the conductive particles, completely giving a totally insulative test result. But insertion as a layer in the material to the right with its insulating matrix (checkerboard) results in a more meaningful measurement. (Two test methods of determining surface and volume resistivity can be averaged to determine shielding effectiveness form 0.1 to 1000 MHz by the graph in **fig. 3**).

To evaluate how well material acts as a shield, let's first recognize that a pure conductive metal performs its shielding effectiveness by an almost 100 percent *reflection of radiation*, whereas a conductive-composite product reflects about 80 percent of radiation and absorbs 20 percent. Mathematically, the shielding effectiveness, expressed in dB, (SE) +R +A or the sum of the energy *A*bsorbed and *R*adiated. In actual practice, there is another term, B, added to A and R, but this term describing internal multiple reflections is usually ignored for materials with greater than 10 dB of SE. This would include just about all materials that are even remotely considered to be shielding elements.

test and measurement systems

There are essentially three methods of measurement: the free-space method, which is impractical at lower frequencies; the shielded box method; and the transmission line technique. The shielded box technique, shown in **fig. 4**, is the most popular. This rectangular metal box, made with thick walls and electrically tight seams in the walls, has a transmitting antenna inside it and a sample port (opening) in one wall for radiation emitted through the port. If the specimen to be tested is placed over the port or hole, and the radiated power P_2 measured against the open power level P_1, SE can be determined by:

$$SE = 10 \log \frac{P_1}{P_2} \text{ (in dB)}$$

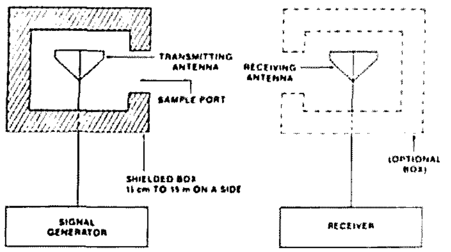


fig. 4. Pictorial illustration of a shielded test chamber.

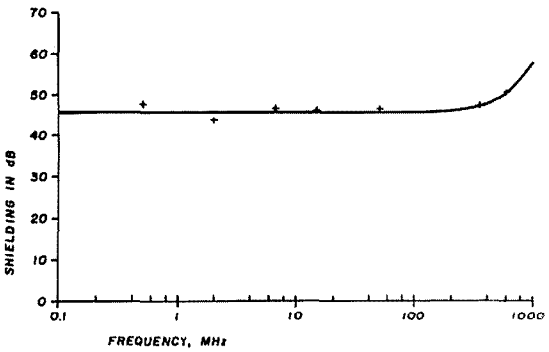


fig. 5. Graph of the dynamic range possible with the coaxial line test technique.

This method of testing has some problems. If there is an edge in the chamber that approaches the wavelength of radiation, resonance can occur with ensuing standing waves between the antennas inside and the chamber's walls. Also, at lower frequencies, the dynamic range is limited to approximately 50 dB because of the energy that can pass through the small sample port at those wavelengths. It is likely that no two boxes would report the same results for a given sample.

The coaxial transmission line technique has no inherent problems associated with it because this method is not frequency-sensitive. (Note that there are no seams in this canister-shaped specimen holder.) Dynamic range, (**fig. 5**), is a function of the generator and the sensitivity of the receiver. Unlike the shielded box, which suffered from the wavelength-to-specimen size ratio changing the chamber's own inherent impedance, this problem does not exist with the transmission line technique.

This method (the TEM or transverse electromagnetic cell) yields reflected, transmitted and absorbed data by reflection coefficient measurements, which allows the designer of the system to optimize reflection. This is ideal for waveguides or parabolic antennas in which the wave is to be redirected.

The main advantage of this measurement technique is the uniformity of repeated test results. The TEM, (**fig. 6**), allows absolute verification to be made that the reflection and transmission ratios in a 50-ohm line are the same as those theoretically obtained in free space on a panel infinite in extent in the plane perpendicular to the incident energy. The dimensions of a 50-ohm line are such that the ratio of the impedance presented by a test specimen transverse to the direction of propagation, to that presented to the free space wave by the same material, is 0.13. This is the exact ratio of 50/377 (the two characteristic impedances).

Electromagnetic fields near an oscillating dipole antenna have a wave impedance as a function of 1/d, where d is the distance between the source and receiver. This is true when the distance to the dipole antenna is very small in comparison to the wave-

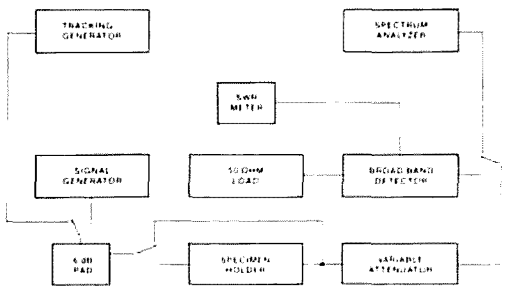


fig. 6. A block diagram of the coaxial line test technique (TEM Cell).

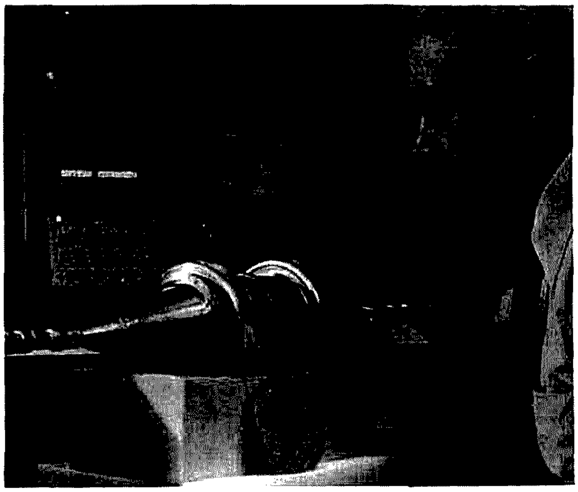


fig. 7. The TEM (Transverse Electromagnetic Cell) test method in operation.

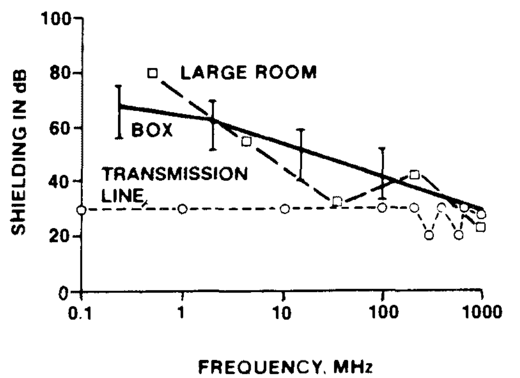


fig. 8. Large room/shielded box/transmission line technique test results.

length and precisely why, at lower frequencies, within a shielded box, the impedance changes with frequency. (We will note some data shortly that graphically reveal this fallacy.) Likewise, as the wave impedance and reflection coefficient increase as frequency decreases, a false indication of an enhanced *SE* results as the frequency decreases.

The coaxial transmission line technique is an insertion-loss measurement technique using a substitution method (see **fig. 7**). The measurements can be made at specific frequencies (the point-to-point mode) using a modulated signal generator, crystal detector, and tuned amplifier (SWR). The other method employs sweeping a band of frequencies with a tracking generator and a spectrum analyzer used as a receiver. The advantage of this point-by-point method is that other information, such as reflection coefficient data, can be taken in this mode; the sweeping mode has the advantage of taking *SE* data at all points, but without the luxury of allowing observation of reflection coefficients.

In the sweeping mode, the signal generator is replaced by a tracking generator and the variable attenuator removed, because the attenuator in the analyzer is used instead. The spectrum analyzer presents the response over the whole bandwidth in a single curve. However, for determination of *SE*, two responses are taken, one with and one without the specimen in the holder. The *SE* is the difference between these two curves.

The dynamic range of this measurement technique is determined by three things: the power level of the signal generator, the sensitivity of the receiver, and the degree of shielding of the equipment and the connecting transmission lines. For example, if a one-watt signal generator with a detector such as an HP415 standing-wave meter are used, an 80-dB dynamic range can be obtained with standard coaxial cables.

Analysis of test results shows how "flat" the test results are from 0.1 to 1,000 MHz, (**figs. 8** and **9**). The difference between measured and predicted results is due to the uncertainty of measurement of the conductivity of the composite plastic, as previously discussed. **Fig. 8** shows a specimen of 20 percent weight piece of Transmet™* series 100 conductivity modifiers in a polyester resin. The methods used in testing were the shielded box method, using a 6.25 foot (2 meter) cubic box in a 31 × 15.5 × 10 foot (10 × 5 × 3 meter) shielded room; the last method was the transmission line technique. **Figs. 10** and **11** show various plastics test results for *SE*.

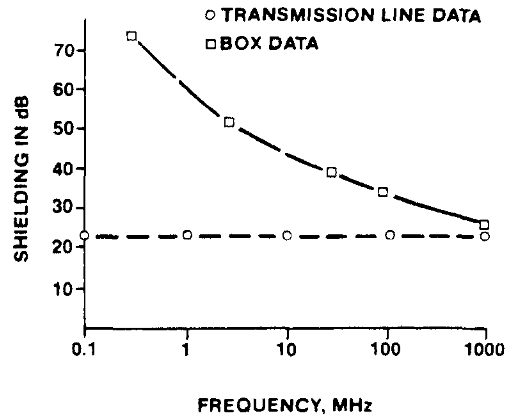


fig. 9. Transmission line technique vs. shielded box test results.

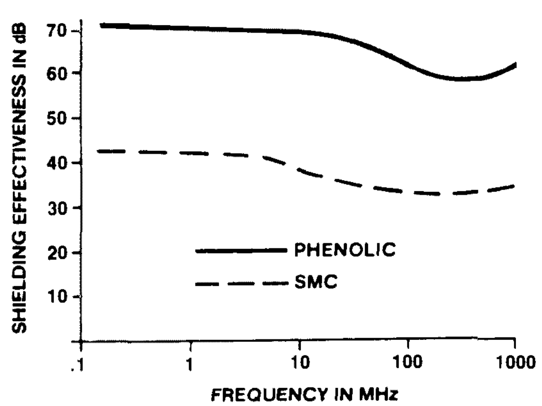


fig. 10. 30-dB SE gained from treated vs. untreated specimens of phenolics.

*The Transmet Corporation, Columbus, Ohio

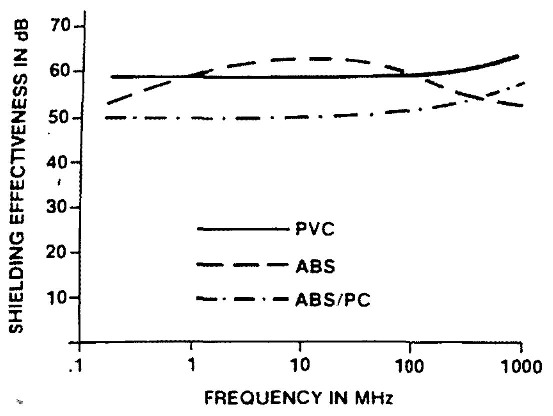


fig. 11. Shielding effectiveness in different types of plastic treated with solidified conductive flakes.

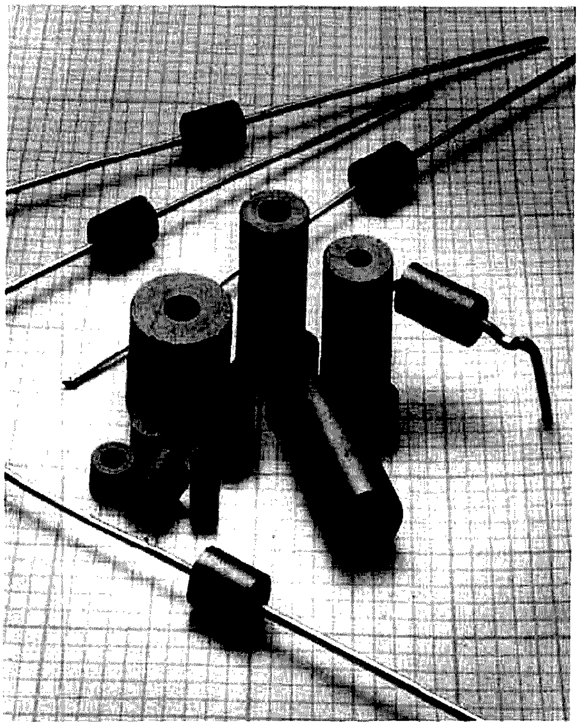


fig. 12. Ferrite beads for EMI suppression.

Although this method of testing is relatively new, it is gaining rapid acceptance. Sometimes called the Stutz method after its inventor, David E. Stutz of the Battelle Institute, this method is used to measure planar waves, whereas the shielded box method is used to measure the near field.

Conducted interference. As stated previously, conducted interference is best attacked at the source or through the cables or wires running to and from the box. These problems are usually of much lower frequency than EMI/RFI emissions and are the result of emissions from motors, transformers, and other magnetic field producers. Most hams know about ferrite beads (**fig. 12**), small ferro-magnetic metal or graphite tubular-shaped devices that slide over wires and suppress conducted emissions. Ferrite beads, however, become saturated at a certain current level; their attenuation does not increase with frequency; they tend to become a resonant frequency element, and there are often gaps because of poor fittings. Thanks to plastics, there are better ways to suppress conducted emissions. Zippertubing™*, for example, uses metallic mesh, aluminum foil laminated to vinyl-impregnated nylon, and high nickel-steel foil in standard jackets to form a protective shield fit over entire cable bundles (**fig. 13**), whether they are the traditional round bunched cables or the flat "ribbon" cables used by printers and modern computer equipment today. A new product now under development by Zippertubing, Inc., uses a plastic polyurethane material, PFR-20B, said to be even more rugged than these other jackets.

Berquist** is another innovative company that makes thermally conductive insulating pads that go between resistors and the PC board. This company, however, has gone one step further by laminating copper right onto the pad, (**fig. 14**). This now provides not only thermal conductivity but also EMI/RFI and conducted interference protection as well. While the metal does cause a capacitance between the transistor and its heatsink, the capacitance is limited to approximately 100 pF for a TO-3 power transistor.

To the ham on a limited budget, the best way to

fig. 13. Zippertubing conductive cable covers for suppression of conducted interference.

*Zippertubing, Inc., Los Angeles, California
**Berquist, Inc., Minneapolis, Minnesota

achieve magnetic shielding might be to use foil tape, which is comparatively inexpensive and comes in thicknesses of less than 0.002 inch to over 0.01 inch thick with permeabilities ranging from as low as 3,000 to over 350,000.

(If you are really fascinated by this subject, you may want to spend $99.50 and buy a kit from the Magnetic Shielding Division of Perfection Mica* (**fig. 15**). This contains sheets and foil forms, a magnetic pick-up probe, design information, and a variety of different types of shielding.)

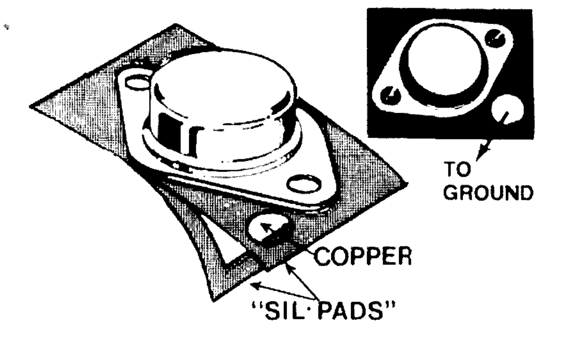


fig. 14. Thermally conductive heatsink pads with built-in conducted interference suppression.

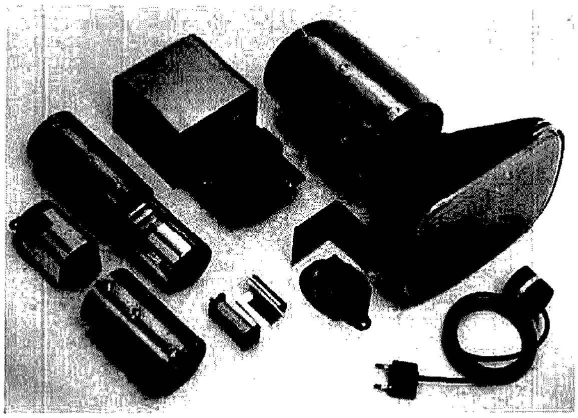


fig. 15. Kit for magnetic field shielding.

There is as yet no perfect coating, adhesive, or potting compound; EMI shielding with plastics is a science in itself. But with technology evolving rapidly in this area — and with many effective EMI-shielding materials becoming increasingly available — you need no longer invariably avoid using plastic cases in building equipment.

ham radio

*Perfection Mica, Inc., Bensenville, Illinois

book and product

REVIEWS

HW-5400 HF transceiver

Regardless of his or her age, every Radio Amateur has heard "the rig here is an HW . . ." Back in the early 1960's, the first of the HW single-banders appeared on the scene, and were an instant hit. That unit spawned a whole line of units which included the SB-101, the SB-301, and many more. There are many more still going strong; and now, there's the HW-5400 microprocessor-controlled HF transceiver.

The HW-5400 has all sorts of interesting features. For instance, when I switched from one band to another, and then back to the first band, it was still tuned to the original frequency. In case of power failure, all the memories are automatically reset to the band edges.

It has split frequency operation, fast/slow AGC, VOX, and antiVOX. It's *all* there. One thing I especially like is the optional keypad frequency entry. I do quite a bit of split frequency work, and this feature makes operating easy; I've also found it handy when going off frequency from a net.

When the kit arrived, the first thing that impressed me was the weight of the power transformer. (Of course, the power supply is rated at 20 amperes, so I shouldn't have been surprised.) As soon as I opened the main carton, I was faced with a notice telling me not to unpack any parts until I was told to in the assembly procedure. I'm glad I followed the instructions. . .the packaging is indeed dense.

preliminary assembly

The assembly of the kit should take about one hundred hours, more or less. It is NOT a kit for someone who has had no experience in building kits. The instructions and illustrations are up to Heathkit standards, but a good deal of care has to be taken during soldering on some of the more tightly packed boards.

I took slightly less than one hundred hours to do the actual assembly, but I did encounter a couple of problems that used up quite a few extra hours. (It should have taken less, but I was being stubborn and didn't want to ask for help.) One of the areas where I got stuck turned out to be caused by a bad FET. And after some discussions with Heath, it turned out that the company was having some difficulties with this particular part. Heath and Motorola, the manufacturer, were working on the problem. The replacement FETs — which were sent promptly — worked fine, so the problem I encountered appears to have been resolved.

The kit has fifteen circuit boards, some of which are simple, with few parts. Others are not. But all are well made, and the silk screening job is good. While there is really no problem in stuffing these boards, I do recommend that you take the advice of good carpenters: measure twice, cut once.

At the end of the assembly of each board there is a series of tests to be made. In one, I found two solder bridges that I hadn't seen in my visual checks — and I was using a lighted inspection magnifier.

I had a great feeling of accomplishment when all the boards were finally done. Then I came to the section innocently titled "Chassis." At that moment I didn't realize that I was about to become an expert in crimping and soldering tiny Molex spring connectors on a multitude of ends of wires. After doing some 50 connections, I checked the price of a crimping tool: over $80. I continued my manual assembly procedures. (Perhaps Heath will develop a design for making a low-cost crimper, because using these connectors really helps during troubleshooting.)

final assembly

As I started the final assembly, I looked at the pile of boards I had just finished, and then at the main chassis, and wondered what kind of magic I was going to have to perform to get all that material onto one tiny chassis. The installation of the pre-assembled main cable harness is a good example of the kind of magic you'll have to perform; I suggest you start this step when you're feeling well-rested and kindly towards the world.

After doing the first two or three steps in the final assembly, I was instructed to mount the audio circuit board and make the necessary connections. Then came the instruction to "connect up a power supply" so that a series of operating tests could be made. It was only then I realized I should have built the power supply first. (I think perhaps it would have been wiser of Heath to mention this at the very beginning.) However, fortunately for me, I had "subcontracted" that part of the job to a ham friend who is normally a very active person (he goes up and down 70-foot towers like I go up and down stairs). He'd just had an extensive operation on one knee, and was "climbing the walls" with inactivity. He built my power supply in about six or seven hours.

All went well with the final assembly until the test of the synthesizer board: I just couldn't get the right readings. After following the "in case of trouble" charts and trying a few tricks of my

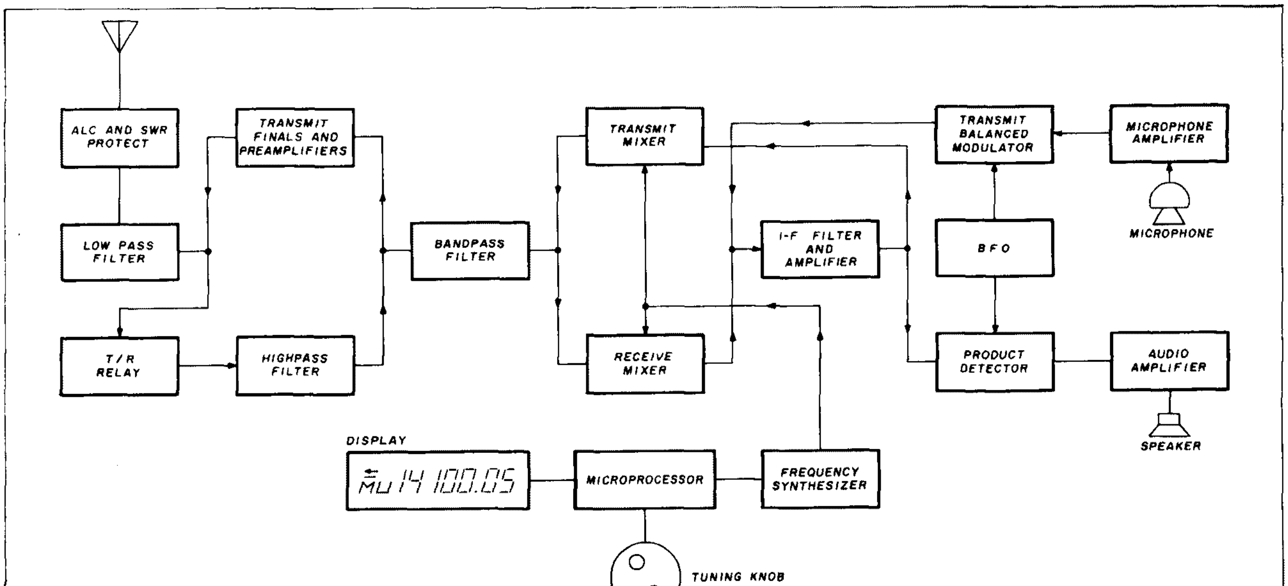


fig. 1. Simplified block diagram of HW-5400.

own, I called Heath. Very shortly, I was talking to Terry, a synthesizer expert, and listening to a quick run-through of that part of the circuit. He made a couple of suggestions and sure enough, his first one identified the trouble. It was a mica tuning capacitor that was marked 100 pF, but was *really* 1000 pF. No wonder the circuit wouldn't tune!

I ran into difficulty again on the IF board, but this time I didn't waste a moment; I got on the phone. It turned out that the trouble was the FET problem mentioned above. A package arrived in three days, and as it turned out, there were four FETs to be replaced.

When I had to remove the IF board, the synthesizer board, and the RF board, I was quite happy that Heath had made me put all those Molex connectors together. Disassembly was easy. It was at this point that the quality of the printed circuit boards became really obvious. I had absolutely no fear of unsoldering and resoldering the connections on the traces of the PCBs.

"smoke" test

There was one interesting thing about doing power-on tests after each step in the assembly. If you've ever put a kit together, then you know the thrill and fear of the "smoke test." With this rig you enjoy the same emotions over and over again. It doesn't lessen with experience.

Now there it was, all put together and ready for use. After three or four contacts on 75 and 20, I decided it was time to finish the job and install the unit in its cabinet. But first, I thought I should do a bit of peaking up, using my old but reliable General Radio Model 605A signal generator. There was no improvement.

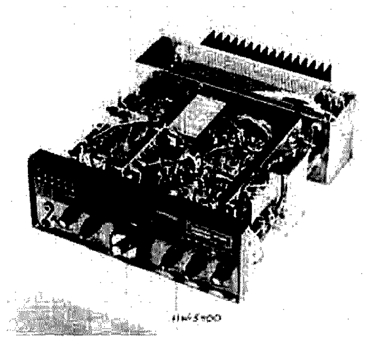
state-of-the-art circuitry

This is a complex, state-of-the-art transceiver. The simplified block diagram (**fig. 1**) shows the flow of signals. On both transmit and receive, the signals go through the lowpass filter (which is switched for each band). On receive, the signal goes through a highpass filter to get rid of broadcast band signals, and then travels through a bandpass filter which is also switched for each band. Notice that the BP filter is used on both transmit and receive. The signals go one way on transmit, and the opposite way on receive. This is done by some very clever diode switching. The final amplifier is permanently connected to the output. The transmit/receive relay is connected in the receive line, and is *opened* on transmit.

finger sensor

Down at the bottom of the block diagram are three blocks labeled DISPLAY, FREQUENCY SYNTHESIZED OSCILLATOR, and MICROPROCESSOR. They are NOT simple. The microprocessor is the device which makes everything happen in the HW-5400. It controls the display, the frequency synthesizer, and the finger sensor of the tuning knob. Yes, that's right . . . the *finger sensor*.

Now here's a bit of genius in design. The main



tuning knob has two finger-spinning holes. One of them has a metal insert. When your finger touches the metal insert, the two digits after the decimal on the display go out, and as you tune, the frequency increment is 1 kHz instead of 50 Hz. So when you want to make a big shift in frequency, use the metal insert spinner; for fine tuning, use the other one. On the unit that I built, I have the keypad accessory; for big jumps I use the keypad, and for 20 or 30 kHz, I use the spinner.

frequency control

The frequency synthesizer circuits control the VCOs, which combine to produce the required injection signal. The VCOs are controlled by a feedback loop referenced back to a crystal controlled oscillator. This loop has a bandpass filter, and the settling time for a VCO is inversely proportional to the bandwidth of the filter. The filter must be narrow enough to get rid of the reference frequency, and yet wide enough to allow quick response.

Heath solves the problem ingeniously. They set up the circuit so that the input to one of the divide-by-N chains is the difference between two VCOs. The step frequency of PLL 1 is 10.05 kHz; that of PLL 2, is 10 kHz. The difference between the two steps is only 50 Hz, yet the reference frequencies are at 10 kHz. This makes the filtering out of the reference frequency relatively easy. The output of VCO 1 is from 5.45-6.05 MHz (in 50 Hz steps). It is mixed with the BFO signal (8.83 MHz) to give 14.28-14.88 MHz. Then the signal is mixed with 10 or 20 MHz, and the output from VCO 3 or VCO 4, and the desired output frequency is produced. A lot of the switching is solid-state, so that the bandswitch is composed of only three wafers. One of those is used to control the solid-state switching circuitry.

The display on the HW-5400 is easy to read. At the left there is a series of symbols that indicate which mode the rig is in. There are several indicators — out of band, split frequency, transmit, and unlocked VCO. The VOX, anti-VOX, delay, and side tone controls are under a little flip-up panel to the right of the display.

When the Heath engineers laid out the printed circuit boards, they kept the transmit and receive parts pretty well separated, and generally speaking, the schematics follow the board layout in the same fashion. This makes it a little easier to locate the components on either the boards or the schematics.

operation

As I mentioned before, when I switched from one band to another, the HW-5400 remembered the frequency I last used on the original band. But it will also remember a second frequency. With eight bands (80, 40, 30, 20, 17, 15, 12, and 10), that makes a total of sixteen frequencies in memory, because the HW-5400 provides for split frequency operation on each band. Incidentally, when I operated split frequency, I found a slight delay in going from receive to transmit (but only on split frequency; simplex, the VCOs are already on frequency and there is no delay). This happened because the VCOs require a little time to settle.

I found it strange to be able to switch from one band to another without having to retune; in fact, I kept looking for the knobs. Fortunately, all of my antennas have fairly low SWR. At least they are low enough so that the HW-5400 will accept them without shutting down. Just to see what would happen, I put the rig on 75 meters and left the 20 meter beam antenna connected. The HW-5400 absolutely refused to put out any power. I didn't do any fancy tests to see at what SWR it would quit, but I figure that this is the mistake most likely to be made.

One disconcerting thing at first was "chirp" of the 50 Hz steps as I tuned in a CW or RTTY signal. But soon I no longer noticed the noise. For final tuning of signals, I started to use the RIT. It is smooth. Another feature I like is the IF shift. It seems to me that whenever I go off a net to talk to someone, the frequency I select invariably becomes the channel adjacent to the National Tuneup Frequency! With the IF shift, I can slide the IF sideways a little and get rid of the QRM. It's a nice feature.

While the high speed tuning feature is great, it did produce a surprise one evening. I spun the dial to a new frequency to meet someone for a QSO, and because Heath included a muting circuit to quiet the receiver while the VCO is settling, I ended up about 15 kHz past a local. He was transmitting when I tuned past him, but it took a few milliseconds before the speaker almost came off the desk from the buckshot. It took me a second or two to realize what had happened. I moved down another 15 or 20 kHz and learned something: the HW-5400 is almost as good as my tube rig for bearing up under close-by locals. I carried on the QSO about 30 kHz from the local, and although I knew he was there, it wasn't difficult or even uncomfortable to carry on the QSO.

Radio Amateurs are never satisfied. It doesn't much matter who designs what . . . we'd always like to see something different. For me, there are two things missing from the HW-5400: 160 meters and a way to connect up my SB-610 band scanner. After having used one for fifteen years or so, I feel almost lost without it connected.

One of the things I really like about this rig is the uncluttered front panel; you can tell what each control is for without searching through the instruction book. The knobs are far enough apart so that I *don't* feel crowded.

If you like to build, and have some experience

at it, the HW-5400 is a good rig to tackle. If you want to dig inside to fix or tweak something, you'll be able to go ahead with confidence because you'll have been inside at least once before. You'll enjoy operating the HW-5400; it's a good rig.

For information on the HW-5400, contact Heath Company, Benton Harbor, Michigan 49022.

Fred Looker, VE3ZL

HW-5400: specifications

coverage:	80,40,30,20,17,15, 12,10 meters plus WWV at 10 MHz
readout:	7-digit display, vacuum fluorescent with special symbols
readout accuracy:	to nearest 50 Hz
frequency control:	synthesized
memory storage:	2 frequencies/band
stability:	less than 50 ppm drift from cold start
modes:	SSB normal and reverse CW — wide and narrow
receiver	
sensitivity:	less than 0.35 μV for 10 dB $\frac{S+N}{N}$
audio output:	2 watts in 4 ohms
AGC:	selectable (fast or slow)
intermod:	70 dB at 25 kHz
transmitter	
RF output:	100 watts except on 10 meters (80 W)
load impedance:	90% rated power at 2:1 SWR high SWR protection
transmit-receive:	SSB: PTT or VOX CW — full break-in (simplex only)
general	
dimensions:	11½ × 14 × 5 inches (29.2 × 35.6 × 12.7 cm)
weight:	24 pounds (10.9 kg)
power:	13.8 volts DC at 20 amperes (HWA-5400-1 power supply: 120/240 V, 60 Hz)

book review: radio frequency design

Wes Hayward, W7ZOI, an engineer from Tektronix, is well known to the Amateur and professional community. His numerous efforts to educate designers in both detailed schematics and system approach have led him to write *Introduction to Radio Frequency Design*, published by Prentice Hall in January, 1982.

Having seen many books on this subject, I was impressed with the way this one was composed and the way issues are addressed. The book consists of eight chapters covering low frequency transistor models, filter basics, coupled resonator filters, transmission lines, two-port networks, amplifiers and mixers, and oscillators and frequency synthesizers. A final chapter is titled "The Receiver: An RF System."

The material presented is complete; references are well chosen. The book reads easily and makes one feel sufficiently impressed to keep on reading. I was tempted to use my computer resources to verify some of the more complex numerical examples, and I was not surprised to find that the author's calculations were correct. I especially enjoyed a few specific items (his non-linear model of the transistor and his analysis of noise, for example) and recommend these sections — and the book as a whole — to all readers. *Introduction to Radio Frequency Design* is thoroughly enjoyable and well worth the $29.95 investment.

DJ2LR

Sherwood filters

Receiver selectivity and the ability to reject unwanted signals are subjects of tremendous interest to Radio Amateurs. Long known for their excellent aftermarket filters and extensive modification of the R4C receiver, Sherwood Engineering now has a line of front end antenna filters. My interest came as a result of a phone call from George Heidelman at Sherwood alerting me to a new filter they had designed recently for the 160 meter DX window, 1.825-1.830 MHz.

If you haven't been on 160 meters recently, you would be surprised by the amount of activity that will be found. On almost any given evening there are plenty of CW and SSB QSO's in progress; rapid growth and the elimination of power restrictions has resulted in some crowding on the band. For instance, SSB operators who sit just above 1.830 MHz with an LSB signal put a fair part of their signal down into the DX window.

During both phone and CW contests, when 160 can sound as bad as 20 meters, the DX window is often bracketed by strong signals. For the average transceiver or receiver, this can result in very difficult copy in the DX window area. Installation of the Sherwood filter will pass only these signals that fall between 1.825 and 1.830 MHz thus facilitating easier reception without front-end overload, intermodulation products, etc. Another suggested application is at large multi-op, multi-transmitter contest stations; a spotter can scan the DX window, minimizing interference, while the operator is working stations on another part of the band. When another multiplier is heard, the operator can QSY to the appropriate frequency and attempt to work the station.

The filters can also be used to reduce interference from high powered shortwave broadcast stations and in high ham density, urban environments. Currently Sherwood Engineering has available filters that cover 25 kHz segments for the 40 and 20 meter CW bands, 12.5 kHz segments of 80/75 meters, and any 5 kHz segment on 160 meters. Other frequencies are available on request.

These low-loss front-end antenna filters are based upon a high performance, 6 pole, crystal 50 ohm design. They have a shape factor of 2.25:1, 6/60 dB and are designed to be used with any receiver and can be adapted to most transceivers.

the acid test

After receiving the filter from Sherwood, I reconfigured my receiver input so that I could switch in and out the filter to do A-B comparisons. Because I live in a low-density area, having other stations nearby is not a problem. However, as more and more new hams have joined the fun on 160 meters, the band has become significantly more crowded. During several 1983 contests, I found that while listening with the filter in, I was able to effectively reduce interference coming from other stations outside the DX window. At the filter's edges, nearby signals were not completely eliminated due to operating at the 3 dB points. My overall impression of these front-end filters is that they would be highly desirable at a multi-op, multi-transmitter operation and in any high ham density area. For 160 meters, they provide an extra measure of selectivity to help ferrett out those weak and hard-to-copy DX signals. They can also be used ahead of a Beverage antenna preamplifier.

Filters for several frequency segments are available for each band. They include the FE-14000/6, FE-14200/6, FE-7000/6, FE-3500/6, and FE-1825/6, as well as others that may be specially ordered. Most are priced at $80; 160 meter filters are priced at $145. Add $3 for domestic shipping, $6 for shipping overseas.

For more information on Sherwood's front-end filter or any of the rest of their product line, contact Sherwood Engineering, 1268 S. Ogden Street, Denver, Colorado 80210.

Circle #302 on Reader Service Card.

— N1ACH

NEW products

remote control antenna tuners

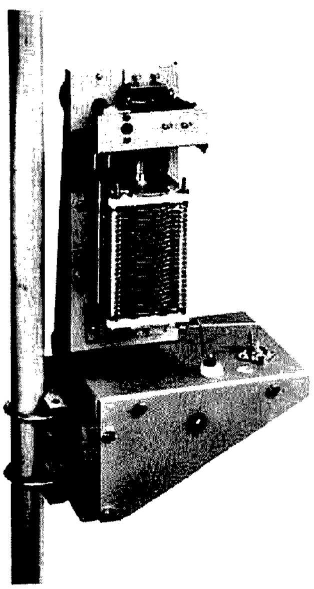
Here's a brand new product of interest to hams who've been disappointed with their antenna's performance. The VT-3/VT-4 is a remotely tuned series-fed capacitor that is connected directly to the antenna. Adjusting the VT-3/VT-4 will tune the antenna to a minimum value of SWR. For example, mobile antennas that would before only tune 20 kHz without readjustment, will now be able to tune the whole band by a flick of the switch. Other suggested uses are with a trapped vertical, half wave doublet, long wire, or sloper.

The VT-3/VT-4 is housed in an aluminum universal base mount with an anodized aluminum cover for weather protection. With the addition of the LC-4 inductor kit, it can be converted to a voltage-fed matching device. This is a single band system that can cover 1.8-30 MHz and will provide full size band coverage with appropriate coil tap adjustment.

The VT-3 is designed for mobile installation and operates directly from the vehicle's 12-volt power supply. The VT-4 is a 117 VAC - 12 VDC supply and switch. The VT-3/VT-4 has a switch to activate the capacitor and limit lights show when either maximum or minimum capacitance values are reached. Both units require eight conductor control cables. Current consumption is low so cable runs can be long.

NEW products

Price for the VT-3/VT-4 is $159.95. Eight conductor cable is available for 24 cents per foot and the LC-4 inductor kit is priced at $39.00. All Vector Radio products are sold with a money-back guarantee.



For more information on the Vector Radio VT-3/VT-4 remote antenna tuner, contact Vector Radio Co., P.O. Box 1166, Cardiff, California 92007.

Circle #303 on Reader Service Card.

radio modem

Macrotronics, Inc. has introduced the RM1000 radio modem, a hardware and software system that converts a personal computer into a state-of-the-art communications terminal supporting Morse, Baudot and ASCII codes. The radio modem is intended for use by Amateur Radio operators and SWL's for copying news and wire sevices. It features commercial quality demodulators, dual bargraph tuning, and extensive software capabilities.

The RM1000 uses multistage active filter demodulators with dual LED bargraph tuning indicators for reception of both Morse code and radioteletype (RTTY) signals. It offers three RTTY shifts which may be selected from the computer keyboard, and a crystal-controlled AFSK tone generator provides stable RTTY keying. Relays are used for Morse code and push-to-talk transmitter keying. A hardware clock continuously displays time and may be inserted into text in a completely user-programmable format.

Many features are included in the software to accommodate a wide variety of operating situations including net operations, MARS, RTTY, art ("PIX"), contesting and SWLing. A 70-page user manual is included.

The RM1000 radio modem system is currently available for ATARI™, APPLE™, IBM™ and Radio Shack TRS80™ microcomputers. For complete information, contact Macrotronics, Inc., 1125 N. Golden State Blvd., Turlock, California 95380.

Circle #304 on Reader Service Card.

new solid-state tube for Drake R-4

Sartori Associates has just announced the availability of a new solid-state tube, the SBA6. Designed to replace the 6BA6 in Drake R-4(A-B-C) receivers, the SBA6 will also replace the RF and IF 6BZ6 vacuum tubes in the R-4A/B and early model R-4C, as well as the 6BA6 and 12BA6 vacuum tubes used in the IF amplifiers of the R-4 series. (For the third mixer in the early model R-4C, we recommend replacing the 6BA6 with a 6HS6/SHS6 — (Drake made this improvement in the mid-model R-4C). The new SBA6 will also serve as a plug-in replacement for your T-4X 12BA6 ALC'd IF amplifier.

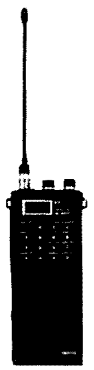
Sartori solid-state tubes provide no-warm-up, high performance, trouble-free operation with R-4(A-B-C) receivers. Sartori also manufactures SEJ7, SHS6, and SBE6 mixers for the R-4 series and the SEJ7, SHS6, SAU6, SAX7-1, SAX7-2, SEV7/SFQ7/SAQ8 for the T-4X series. All are priced at $23.00 postpaid.

For more information about the SBA6, contact Sartori Engineering, P.O. Box 2083, Richardson, Texas 75080.

Circle #305 on Reader Service Card.

IC-04A and IC-04AT

The IC-04A and IC-04AT, two new 440 MHz HTs from ICOM, feature frequency entry, control functions and 32 PL tones controlled by the 16-button pad on the face of the radio. Also in-



cluded are priority, scanning (both of memories and programmable band scan), and DTMF (04AT only). For scanning, 5 kHz increments are front-panel selectable. Ten memories with internal lithium battery backup give the ultimate in flexibility for channelizing operation for easy access to most-used channels. A custom LCD readout with S-meter is unique to the ham industry.

The IC-04A and IC-04AT have the same styling, control features and functions of the IC-02A(T), and utilize the existing accessory line available for the IC-2A and IC-2AT, plus new accessories such as long-life and high-power battery packs.

For details, contact ICOM, 2112 116th Avenue, N.E., Bellvue, Washington 98004.

Circle #308 on Reader Service Card.

full-function DTMF decoder

A full-function, dual-tone multifrequency decoder module, model 2009, is a state-of-the-art CMOS design which decodes all 16 DTMF codes. Available from Proham Electronics, model 2009 has several advanced features such as a crystal-controlled timebase for long-term accuracy, on-board voltage regulation, counter detection with period averaging to minimize falsing, and latched 4-bit digital outputs with a choice of binary or row/column format.

Two LSI CMOS chips provide high performance operation and minimize the parts count. All bandpass and band reject filtering is achieved by using one switched-capacitive filter integrated circuit. The time base oscillator and dividers are also within this chip. Likewise, the actual decoding is performed by the second CMOS device. This simplifies operation, since there is only one adjustment required, and it sets the operating level. There are no frequency adjustments because all timing is referenced to the crystal controlled time base. Ancillary benefits resulting from the application of these LSI CMOS devices is compact size, the printed circuit board is only 3.6 × 2.0 inches, and low power requirement, typically 40 mA at 12.5 VDC. The kit is easy to

build using the comprehensive instruction manual supplied, and easy to use. When driven with an audio signal beween 50 mV and 1.0 V, the model 2009 produces a 4-bit digital output code corresponding to the DTMF digit detected and valid code pulse. This output can be used to drive a parallel port of a microcomputer or additional digital logic circuitry as required. The price of a bare board with manual is $9.95; board, manual, filter and decoder chip, $44.95; complete kit, $99.95; and manual only, $5.00. All prices include postage in U.S.A. (Ohio residents add 5% sales tax.)

For further information, contact Proham Electronics, Incorporated, 34620 Lakeland Blvd., Eastlake, Ohio 44094.

Circle #306 on Reader Service Card.

shirt-pocket volt-ohmmeter

The new Model 3525 DIGI-PROBE™ volt-ohmmeter, just introduced by Triplett, is said to be one of the smallest trouble-shooting, battery-operated, digital instruments presently manufactured. Its shirt-pocket size, internal overload protection, accuracy and auto-ranging features make it appropriate for a myriad of

lab, circuit design or in-field measurements on industrial, commercial, or consumer electronic/electrical equipment.

Only 6-3/8″ long × 1-1/8″ wide × 3/4″ deep (162 × 28 × 20 mm), the DIGI-PROBE™ utilizes a large 5mm easy-reading 3.5 digit LCD display with a convenient "Data Hold" feature to facilitate measurements in low ambient light or in confined areas. It also enables the user to "hold" the reading for later review. An instant-tone continuity test permits rapid testing of diodes, shorts and circuit continuity. Volts, ohms, and continuity are easily selected with a simple function switch. AC and DC volts are selected by push-button with AC shown on the LCD display.

Auto-ranging on volts and ohms functions eliminates the need for range selection, providing true "Touch and Test" capability. The Model 3525 has thirteen ranges. Range selection in all functions is fully atuomatic. The ranges are: 0-500 VDC in four ranges; 0-500 VAC in four ranges; 0-2.0 Megohms (2.9 Megohms in overrange) in four ranges. Auto-ranging response time is 5 seconds maximum and accuracy is ±0.75 percent of RDG + two digits on most ranges. Blinking-digit overrange indication and low battery visual indication are provided. Internal overload protection is to 750 VAC/DC in voltage ranges and 250 VAC/DC in ohms and continuity ranges.

The DIGI-PROBE™ case is molded of high impact black thermoplastic with textured surface. The unit weighs only 2-1/2 ounces (0.07 kg).

Priced at $65.00, the DIGI-PROBE™ is furnished with two 1-1/2 volt button-type batteries, shirt-pocket carrying case, attached 28-inch test lead, comprehensive instruction manual and one-year warranty.

For information or a free demonstration of the DIGI-PROBE™ contact Triplett Corporation, One Triplett Drive, Bluffton, Ohio 45817.

Circle #309 on Reader Service Card.

noise figure measurements application note

HP's new application note AN/57-1, "Principles of RF and Microwave Noise Figure Measurement," is now available for all those working on device, component, sub-system and system noise figure. It replaces the long-popular AN/57, *"Noise Figure Primer."*

The 40-page note serves as a comprehensive tutorial on noise figure, with detailed material on thermal and shot noise, concepts of noise figure, effective noise temperature, Y-factor, etc.

Plenty of useful information is provided on subtle measurement considerations including single sideband vs. double sideband, effects of local-oscillator noise, second-stage effects and corrections, hot/cold techniques, frequency conversation and image considerations.

An extensive glossary includes common symbols and detailed technical explanations of most terms. Also included is a bibliography of 34 other noise-figure-related references.

For a free copy of AN/57-1, contact Hewlett-Packard Company, 1820 Embarcadero Road, Palo Alto, California 94303.

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Coming Events

ACTIVITIES

"Places to go..."

OHIO: The Cuyahoga Falls ARC's 30th annual Electronic Equipment Auction and Hamfest, Sunday, February 26, North High School, Akron. 8 AM to 4 PM. Tickets $2.50 advance, $3.00 at door. Sellers may bring own tables or some available for $2.00, advance table reservations advised. Talk in on 87/27. For information: CFARC, P.O. Box 6, Cuyahoga Falls, OH 44222 or call K8JSL (216) 923-3830.

INDIANA: The LaPorte Amateur Radio Club's Winter Hamfest, Sunday, February 26, Civic Auditorium in LaPorte. Starts 7 AM Chicago time. Admission $2.50 per person. 8 ft. long tables available for $2.00 each by reservation. Good food, coffee, etc. Talk in on 52 simplex. SASE for tables, tickets, or information to LPARC, P.O. Box 30, LaPorte, IN 46350.

MICHIGAN: The 14th annual Livonia Amateur Radio Club's Swap 'n Shop, Sunday, March 4, 8 AM to 4 PM, Churchill High School in Livonia. Plenty of tables, refreshments and free parking. Talk in on 144.75/5.35 and 52 simplex. For further information send large SASE to Neil Coffin, WA8GWL, Livonia ARC, P.O. Box 2111, Livonia, MI 48151.

OHIO: Cincinnati ARRL '84 State Convention and Flea Market, February 25 and 26. Registration $5. Flea market $4 per space both days. Forums, meetings, vendors, Wouff Hong, banquet. Hospitality suite Friday and Saturday nights. Write: Cincinnati ARRL '84, POB11300, Cincinnati, OH 45211 or call (513) 825-8234.

KENTUCKY: Glasgow Swapfest, Saturday, February 25, 8 AM CST til . . . Glasgow Flea Market Building 2 miles south of Glas-

gow on highway 31E. Large flea market, free coffee, free parking. Admission $2 per person, no extra charge for exhibitors. One free table per exhibitor, extra tables available for $3.00 each. Talk in on 146.34/94 or 147.63/03. For information: WA4JZO, 121 Adairland Ct., Glasgow, KY 42141

NEW JERSEY: Springfest '84 sponsored by the Shore Points ARC, Saturday, March 10 from 9 AM to 4 PM, Atlantic County 4-H Center, Egg Harbor City, approx. 15 miles west of Atlantic City. Large heated, indoor selling space. Covered tailgating (weather permitting). Sellers $5 per space with own table. Buyers $2.50 advance, $3.00 day of hamfest. For information, SPARC, P.O. Box 142, Absecon, NJ 08201.

PENNSYLVANIA: The 1984 Lancaster Hamfest, Sunday, February 19, Guernsey Sales Pavillion, U.S. Rts. 30 and 896, Lancaster. 0800 to 1600, dealer setup 0600. Commercial tables (main hall) $15.00. Non-commercial (rear annex) $6.00. General admission $3.00. Tailgating free with general admission (weather permitting). Talk in on 146.61 and 147.015. Send reservations to Hamfest Committee, P.O. Box 6082, Lancaster, PA 17603. Please make checks payable to SERCOM, Inc.

MICHIGAN: The Cherryland Amateur Radio Club's 11th annual Swap N Shop, February 11, Immaculate Conception Elementary School gym, 218 Vine Street, Traverse City, 8 AM to 2:30 PM. Tables $3.00 each with setup at 6:30 AM. Admission $2.50. Talk in on 146.25/.85. For details SASE to Jeny Cermak, K8YVU, 3905 Slusher Rd., Traverse City, MI 49684

OPERATING EVENTS
"Things to do..."

FEBRUARY 1 TO 12: NORAC (North Okanagan Radio Amateur Club) will be operating a special event station during Western Canada's largest Winter Carnival. NORAC will operate VE7NOR from 21.00 Z to 24.30 Z during this time in the General portion of each band about 50 KC up. Call "CQ Winter Carnival Award". The award is available to all Amateurs Worldwide who contact 3 Vernon area stations or one QSO with club station. The award is free but $1.00 or 2 IRC's to cover postage would be appreciated.

MARCH 1984 / $2.50

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contents

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REFLECTIONS

Meeting the Challenge of Volunteer Testing

The VEC program is now a reality — or it will be by the time this issue arrives in your mailbox. While it's not yet in place everywhere, proposals for volunteer exam programs have already been submitted to the FCC for at least three of the thirteen areas identified as the responsibility of a specific regional volunteer examiner coordinator (see *Presstop,* page 8). In addition, proposals for several other areas are expected in Washington very soon. Indeed, it appears likely that much of the country will again be enjoying regular and frequent opportunities for Amateur license exams before spring is very far along.

This is good news, because access to the examination process is essential to the survival of Amateur Radio. As of January 1, 1984, even major FCC field offices that had offered weekly examinations to walk-in applicants were cut back to only four one-week examination periods a year, with tests now scheduled by appointment only. Would-be Amateurs and those anxious to upgrade who live near the smaller FCC offices have even less opportunity.

Before January 1, Amateurs were taking exams at a rate of about 2000 a month. And even though a significant number of these were Novices (who were already being tested by Amateur volunteers), the fact remains that trying to cram all the rest of the applicants into a drastically reduced FCC-administered program would inevitably result in chaos. That's why it's so encouraging to find the Amateur community eager and — at least in some places — ready to take up the challenge of volunteer examination. All that's needed now is someone to accept that challenge in those areas as yet unclaimed.

What does it take to get a functioning VEC going in your area? Must a VEC be a part of a large, national organization with a full-time staff devoted only to Amateur Radio? *No.* Let's take a look at who has already seized the opportunity or who seems to be moving rapidly in that direction. One is a large but regional radio club (Alaska); another, a major repeater group (New York/New Jersey). There's a MARS organization taking responsibility for testing in the Virgin Islands and Puerto Rico; a local radio club that also sponsors the world's largest hamfest (8th call area); and an educational institution (9th call area). A diverse group, you'll notice, with one thing in common: a membership composed of dedicated, concerned Amateurs who saw the need and made the substantial effort necessary to get their organization involved!

Your area has many potential VECs. You can start with a strong local radio club, hamfest sponsor, or repeater group — or better yet, a regional Amateur organization such as a state or area repeater or radio club council. You might even seek support from outside Amateur Radio, from a school or business that has enough Amateurs to sponsor its own Amateur Radio club. Their corporate interest can be easily justified, since the VEC will perform a valuable public service, one that will earn the recognition and appreciation of the Amateur community and — if the Grenada experience is fresh in the public mind — of the larger community as well.

Once you've found your potential VEC, the first thing you'll need to do is get a copy of the FCC's Report and Order on PR Docket 83-27. It can be found in Volume 48, No. 195 — October 6, 1983 — of the Federal Register, available at many libraries and law offices; we'll also send you a copy for an SASE. (Enclose 71¢ postage.) Study it thoroughly, so you'll know precisely what responsibilities you're getting your group or organization into. If you're asking a school or corporation to become a VEC through its radio club, prepare a pursuasive sales pitch on the potential benefits of the program: public service, goodwill of present and future students, employees, neighbors, etc. Don't forget to tell them that their out-of-pocket expenses are (or soon will be) reimbursable up to $4 per exam; make it clear that you're asking for support, not charity.

After you've reached agreement with your future VEC, all that's left is making your VEC proposal to the FCC. Here's where careful study of that FCC Report and Order will pay off. If you've done your homework, you won't weaken your proposal by asking to do something the FCC has specifically forbidden. If you're really smart, you won't propose doing *anything* more than you need to do. Why? Because proposing more than delivery of the service required may lock you into something that may turn out to be impractical, unnecesssary, or simply more than your group can handle efficiently. The FCC knows that. All they really want from you is a reasonable, well-developed and clearly thought out proposal from a competent, qualified organization. Give them that, and you'll get the job!

The volunteer examination program may be the biggest and most important administrative challenge that Amateur Radio has ever faced. Let's show the FCC that its faith in Amateur Radio is well placed!

Joe Schroeder, W9JUV
Associate Editor

As a service to readers, *ham radio* will publish a continuously updated monthly listing of all VECs. We'll need your cooperation to do this; if you are affiliated with a group that has received VEC status, or is seeking approval as a VEC, please let us know.

Rich Rosen, K2RR
Editor-in-Chief

presstop de W9JUV

THREE REGIONAL VOLUNTEER EXAMINER COORDINATORS SHOULD HAVE FCC APPROVAL by the time this reaches print, with several more close behind. First in with a satisfactory proposal was the Anchorage Radio Club, which will oversee Amateur exams for Alaska. Metroplex, a New York repeater group with over 1000 members, has volunteered to perform the VEC task in New York and New Jersey, and a MARS group in Puerto Rico has asked to serve the Caribbean. In the ninth call area DeVry, a highly respected technical school which does not teach Amateur Radio but has a very active Amateur club, has proposed not only acting as the VEC but also offering Amateur exams weekday evenings and Saturdays on its Chicago campus.

Additional Proposals Are Also Expected From the Dayton Amateur Radio Association, the Hamvention sponsor, for the eighth call area, while a Texas group is believed preparing to ask for the fifth district responsibility.

The Reimbursement Of VEC Expenses Is Still In The Future, waiting for the necessary change to be incorporated in Part 97. Though the ARRL filed a Request for Agency Action to have the $4 fee go into effect immediately, the Commission feels that justification for and accounting of the fees collected is needed. It appears likely that they'll want to have appropriate procedures and guidelines developed in response to a rule making procedure.

THE FORMER WB6JAC WENT TO JAIL JANUARY 23 to begin serving an 18 month sentence for operating without a license. Richard A. Burton, who had earlier lost his license for transmitting obscene language, was back in court on four charges that he'd continued to operate after the FCC lifted his license. In his second appearance before U.S. District Court Judge Manuel Real, Burton was sentenced to 12 months in federal prison followed by six more months in "a jail type of facility." In addition, Judge Real also imposed five years of probation on Burton, during which he "shall not be found in any place in which any kind of broadcast is made by radio or otherwise."

Burton Was Originally Indicted in May, 1982, both for operating without a license and for transmitting indecent language. His license had been revoked earlier for his indecent transmissions. He was convicted on all but one count, though the convictions for indecent language were later thrown out by an appeals court. It appears Burton will be the first person to go to prison for violation of the Amateur Radio rules since the late 1940s.

In An Unrelated Case, Out-Of-Band Operation Has Cost two East Coast Amateurs healthy fines. KA2QMX and KA2GWV/4 were both caught operating outside the 40-meter band using false callsigns. Their repeated violations cost KA2QMX, a Technician class who refused an FCC request to inspect his station, $1100, while KA2GWV paid a $600 "monetary forfeiture." Such out-of-band operation is not uncommon, but only rarely practiced by licensed Amateurs.

ACTION ON THE VOLUNTEER ENFORCEMENT PROGRAM, at a standstill since last fall, may see some action soon. A draft of a "prototype" monitoring and enforcement training manual, developed under a joint FCC/ARRL effort, has been awaiting approval by the League Board of Directors since late last fall. However, it seems likely that the Commission, feeling the need to get some form of direct Amateur involvement in rules enforcement underway, will be proceeding soon, with or without any formal League sanction.

PUBLICATION OF W5LFL'S STS-9 LOG DISAPPOINTED MANY would-be space DXers. The third and ninth call areas fared the worst, with just a few QSOs each, while the fours, fives, and sixes dominated the roster of fortunates by logging three dozen or more contacts each. Outside the U.S., Canada (including VO) made the log 18 times, almost 40 European stations (including 11 Germans), nine VKs, eight from all of Latin America, and only one (JY1) from Asia; no one from Africa made the list. Further logging tape work could pull out a few more calls, but shouldn't change the proportions appreciably.

The Excellent PR From W5LFL's Space Operation and other Amateur Radio/media topics will be the subject of a Dayton Hamvention panel discussion. Joining moderator Jim Davis, KU8R, will be Steve Mendelsohn, WA2DHF (CBS), Bill Pasternak, WA6ITF (Metromedia and Westlink/Westlink Report), Bob Hanson, W9AIF (Grey-North/Electra), Pete O'Dell, KB1F (ARRL, and Joe Schroeder, W9JUV (ham radio).

NOMINATIONS FOR DAYTON HAMVENTION'S "HAM OF THE YEAR," "Special Achievement," and the new "Technical Excellence" Awards are due by the end of March. Nomination letters go to the Awards Committee, Dayton Hamvention, Box 44, Dayton, Ohio 45401.

Europe's Hamvention Equivalent, The "Ham Radio" Convention held in Friedrichshafen, Germany, expects over 10,000 Amateurs from throughout Europe June 22-24. Amateurs who'll be in Europe then can get details from the DARC, Box 1155, D3507 Vaunatal, West Germany.

ADDITIONAL FREQUENCIES FOR RACES OPERATION in the event of a major national emergency have been authorized by the FCC. Acting on PR Docket 83-524, the Commission added all the frequencies originally proposed plus both the old and new 2-meter repeater subbands to the frequencies that would be available if the President invoked his emergency war powers.

Further Expansion Of The HF Amateur Phone Bands is not likely in the near future. FCC staff limitations combined with decreasing MUFs and resultant lessened activity on 28 MHz will probably push any changes on that band off indefinitely.

6-METER ACTIVITY FROM EUROPE WILL BE INCREASING sharply in the near future. The RSGB reports the British government plans to increase special 50-MHz operating permits from the present 40 to 100 this year. Applications must be submitted before March 31.

comments

compact SSB

Dear HR:

I enjoyed Rick Littlefield's article on the compact SSB receiver (November, 1983). However, I do not agree with the use of No. 2 mix toroid cores for VFO use. No. 6 mix cores are much more temperature-stable than No. 2 mix cores. The most stable VFOs for HF oscillators are constructed with No. 6 mix toroid cores and NPO capacitors. Perhaps I am being a little picky about this, but using the right core material for oscillators can make a big difference in short and long-term drift.

Paul Montgomery, KA0GPE
Westcliffe, Colorado

phased verticals

Dear HR:

When I tried to design a 2-element vertical phased array, I ran into two problems. Where, electrically, did I feed the array and what was the velocity factor (VF) of my RG58U? Not sure what to do, I filed the matter.

Then good fortune appeared in the form of Forrest Gehrke's series of articles on the subject in *ham radio* (May, June, July, October, December, 1983). I have followed the series with great interest using an HP29C and a Smith Chart. The ABCD parameter analysis in the last article tied it all neatly together for me. I was more sure of what to do and why.

What to do with all this information?

1. In a narrow east-west back yard, I want to set up a 20-meter 2-element array cut for the low end of the CW band — and point it toward Quebec.
2. I know a 2-element Yagi would work, as I made one out of wood, aluminum, and old fence post insulators in 1959. I want to experiment.
3. In the spring, if the temperature ever gets above zero, I intend to put up the version of the array K2BT discusses in his October article. This is the array with the quarter-wavelength feeders and the 90-degree delay line.
4. Then I'll measure Z at the receiver with a noise bridge. Knowing the VF of my coax, I'll move with a Smith Chart toward the common feeder point — and see if K2BT's values are correct.

I'm weak on antenna theory, and for a change it's nice to know precisely where and why.

David Winter, W9OAM
Amboy, Illinois

short circuit

good news department: auto dialer

Readers who wish to construct the state-of-the-art auto dialer described in K2MWU's December, 1983 article may order the auto dialer in kit form from Tek-mation, Inc., 2618 North Stowell, Milwaukee, Wisconsin 53211, for \$34.95 plus \$2.50 postage and handling. (Note that the kit does not include the MD-22 chip; this is available from CES, Inc., P.O. Box 2930, Winter Park, Florida 32790.)

Printed circuit boards for the auto dialer are no longer available from Dynaclad, but may be ordered from Tek-mation or from Circuit Board Specialists, P.O. Box 969, Pueblo, Colorado 81002. The price per board — from either supplier — is \$7.50 plus \$1.50 postage and handling.

QSL cards

If you're still waiting for your free QSL cards from RCA (see August, 1983 issue, page 23), relax — they're on the way. Over half a million cards have been printed and shipped; more have been ordered to fill the remaining requests. — **Editor**

build your own "audio to microwave" amplifier

Readily available modular hybrid amplifiers provide extreme bandwidth, exceptional performance

How many times have you tried to build a wideband amplifier to cover the VHF or UHF bands only to find that you've constructed not a wideband amplifier, but a very poor oscillator or the world's most nonlinear active attenuator? At those frequencies, both careful design and close attention to physical configuration are mandatory. Individual component response must also be well characterized. The active devices must be particularly pampered to assure stability over all operating conditions. Although the design of input and output matching to assure stability is relatively straightforward, few of us have the necessary experience to make this a painless exercise with consistent results. It is here that the modular hybrid amplifiers provide the more critical elements of interfacing to the active devices.

In this article we will briefly review modular RF amplifiers with particular emphasis on some of the more critical aspects of their use. We will then go through the step-by-step construction of a modest RF amplifier using these components, which will perform quite well from audio into L-band. The devices specifically discussed are those with which I have had personal experience; many other manufacturers produce similar components that are equally as suitable.

the modular hybrid amplifier

Basically, the modular amplifier consists of one or more high-frequency active devices (usually bipolar transistors), mounted on a hybrid substrate with other discrete and deposited components. This completed substrate, which resembles a miniature PC board, is mounted in some type of header (a TO-12, for example) to provide a controlled physical environment as well as a simple means for connecting to the device. (A typical device is shown in fig. 1.) These devices are available in numerous circuit configurations, but in general they are all simple feedback amplifiers. Feedback helps stabilize the amplifier parameters and make them less sensitive to the characteristics of the active devices. This in turn helps lower cost and provides more consistent performance.

The simplest configuration is that of the devices similar to the Motorola MWA series of amplifiers. A simplified schematic of the MWA configuration is shown in **fig. 2**. Deceivingly simple, isn't it? This configuration requires the addition of input and output coupling capacitors, a collector load resistor, and a power supply decoupling capacitor. The external coupling capacitors allow the user to set the low frequency cutoff where desired. However, as the lower cutoff frequency is reduced, the physical size of the coupling capacitors increases, generally causing poorer RF performance. A practical limit is on the order of a few hundred Hertz. An external collector resistor allows tailoring the circuit to a wide range of supply voltages. The inductor L_C is part of the internal matching network. The MWA-type components are available with upper cutoff frequencies greater than 1 GHz.

A second popular configuration is that similar to the Avantek GPD 400 and GPD 1000 series of devices. A simplified schematic of the GPD 400 device is shown in **fig. 3**. These are very similar to the MWA type devices with the inclusion of the collector resistor and coupling capacitors. Since the collector resistor is fixed, these are basically fixed supply devices. Also, the coupling capacitors are reasonably small due to limited space in the package. This limits the lower cutoff to about 5 MHz. However, the GPD 400 and GPD 1000 series components are available without internal coupling capacitors. This allows the user to set the lower cutoff at almost any frequency within reason. As with the MWA devices, a few hundred Hertz

By Michael E. Gruchalla, 2450 Alamo Avenue, S.E., P.O. Box 9100, Albuquerque, New Mexico 87119

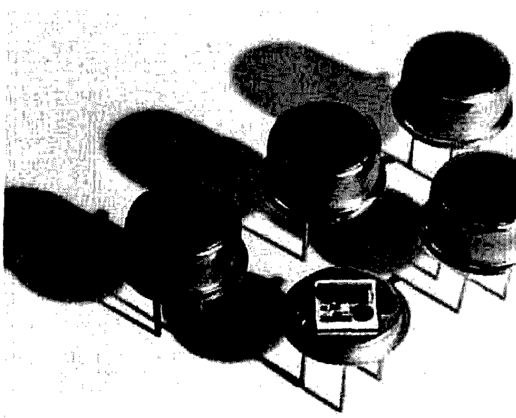


fig. 1. Hybrid modules are the basic building blocks around which the broadband amplifier is constructed.

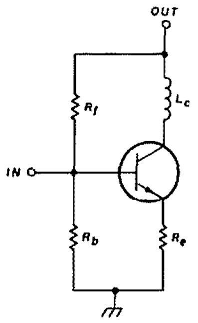


fig. 2. Simplified schematic of a 100 MHz bandwidth RF amplifier designed around the MWA module.

is a reasonable practical limit. Also, elimination of the internal coupling capacitors reduces the cost by about a factor of three. Components of the GPD configuration are available with cutoff frequencies exceeding 1 GHz.

Another configuration is that of the Avantek UTO devices (**fig. 4**).[1] Although it may not be obvious at first glance, these are very similar to the GPD 400/1000 configurations. The base resistor has been eliminated and a capacitor has been added in series with R_f. Next, a voltage regulator, U1, has been added to accurately control the operating point of Q1. This, in turn, offers more stable overall operation. These, however, may have a price more than a factor of 10 above a GPD type part with similar RF specifications. This component is useful in exacting applications with the more modest GPD type part serving very well in the majority of cases. Unlike the GPD parts, the UTO series is not available without the internal coupling capacitors. This is due to the added decoupling needed in the voltage regulator (two more pins would be needed on the package to allow for the addition of external capacitors). The lower cutoff is limited to 5 - 10 MHz. Upper cutoff frequencies as high as **2300 MHz** are available.

Finally, there are several configurations using transformer type feedback networks. This general configuration ideally has the potential for lossless, noiseless feedback. A typical configuration using a directional coupler is shown in **fig. 5**. The principal advantages of this feedback configuration over simple resistive feedback are the capability of providing lower noise figure and delivering higher power to the load. One disadvantage of this configuration is that the bandwidth is limited to about two or three *decades*. In the case of modular amplifiers, the lower cutoff is generally in the vicinity of 5 to 10 MHz because of the limited size of the transformers. Also, these units tend to be more expensive than simple resistive feedback devices because of their added complexity.

noise figure

A carefully designed and fabricated amplifier system using hybrid modular amplifiers can provide a noise figure on the order of 4 to 6 dB (about 400 degrees K to 900 degrees K). This is certainly not in the same class with typical LNA's (low noise amplifiers — typically 120 degrees K — Editor), but it is nevertheless reasonably good performance for an inexpensive general purpose amplifier. The total finished cost of the amplifier described below is perhaps a factor of two or three below the typical cost of just the input active device of an LNA! In many applications, a nominal 5 dB noise figure is quite adequate. Only the most exacting applications justify the cost associated with lower noise figures. Also, the modular amplifiers take all of the guesswork and "tweaking" out of amplifier construction. Few of us have the necessary equipment to accurately measure noise figure, particularly below 5 dB or so. Carefully designed and constructed amplifiers using the modular hybrids will yield very predictable and consistent performance. The noise figure will be at least that specified for the first stage amplifier and will generally be a little better if proper care is taken in the selection of components and construction.

The noise figure of the overall amplifier can, however, be compromised severely by the use of poor resistors in the amplifier assembly. Some resistors exhibit considerable excess noise. This noise, introduced particularly into the first stage, as well as all other stages, could increase the noise figure by as much as 5 dB. Also, a type of "popcorn" noise, or random spikes, often occurs in these resistors. The energy of the individual spikes is low, yielding an RMS component which gives the 5 dB noise figure degradation mentioned. However, the peaks of the spikes are quite high — perhaps a factor of 10 or 100 higher than the RMS level. If the noise figure is computed in a way that uses the peak amplitudes of these pulses, a noise

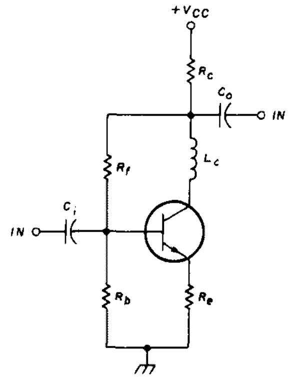


fig. 3. Simplified schematic of a wide bandwidth RF amplifier designed around the Avantek GPD series modules.

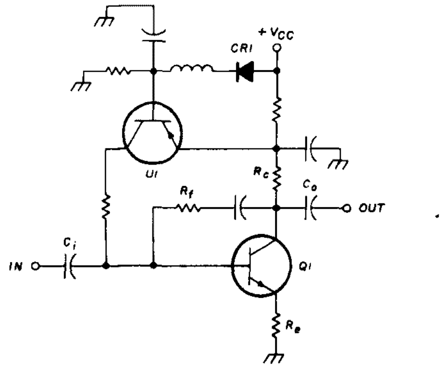


fig. 4. Schematic of a broadband amplifier that incorporates the Avantek UTO series of modules. More precise and stable operation is associated with slightly greater circuit complexity.

figure of 20 dB to 40 dB could easily be achieved. Such a computation, however, is not a correct mathematical exercise because a specific value of a noise spike cannot be computed, and specifying a "peak noise level" is generally meaningless. This is why noise is specified in terms of its RMS value (a statistical term) or its mean square value (square of the RMS value) proportional to noise power. So, the "popcorn" noise may exhibit a reasonably low mean square noise, while still having large amplitude spikes. Viewing this type of noise on an oscilloscope, one would see a low noise floor with high narrow spikes. If the displayed noise were set at about 0.1 division, the spikes could be as high as 1 to 10 divisions. These impulses exhibit a very wide spectrum and can destroy the usefulness of the amplifier at all frequencies. The best solution to this problem is to use high quality, name brand resistors; avoid the very inexpensive carbon film types. (An excellent review of noise considerations is available for those who wish to delve deeper into this subject.)[1]

do's and don'ts

Although the modular hybrid units are reasonably straightforward to use, there are a few do's and don'ts that are important to consider in order to avoid problems. The two most important considerations are proper mechanical mounting and good high-frequency bypassing of the power supply. Also, the use of a groundplane is an absolute *must.*

The mechanical mounting of the device must assure that the case is well grounded electrically. It must also provide good mechanical contact with the groundplane for proper heatsinking. Some manufacturers such as Avantek provide a mounting kit (**fig. 6**). Use it! A word of caution, though — you must mechanically mount the component *before* soldering its leads. If you solder the leads and do not have the package firmly against the groundplane, you may break the lead penetrations into the package and cause internal damage to the part when you tighten the mounting kit. When no mounting kit is available, it is a good practice to tack solder the flange of the package to the groundplane in two or three places. The index tab is an excellent place for one of these solder connections.

I have found one especially useful variation to the manufacturer's mounting. This concerns the ground lead. Normally, a plated through-hole from the groundplane to a trace-side pad would be provided for this lead. If the part is mounted in this manner, it is difficult to unsolder and remove the part without damaging the board, the part, or both, due to the heatsinking action of the groundplane. If the groundplane is relieved around the ground lead, as with the other leads, the removal problem is eliminated. However, a good electrical ground must still be provided. Good mechanical mounting will generally provide good electrical grounding, but to be sure, a pad configuration similar to that shown in **fig. 7** should be used. Adequate grounding is provided by the triangular ground pad. The two through-holes should either be plated through to the groundplane or should have short pieces of bus wire soldered in as feedthrough conductors. One screw of the mounting kit also ties to this pad, providing an additional ground path. The short conductors between the V_{CC} pin and ground pin provide a place to mount a high-frequency chip capacitor for V_{CC} bypassing with minimum inductance. I have found that up to 1 GHz and somewhat higher, this grounding configuration works quite well. Now, using this mounting technique, when you dig through your junk box for an old board with a good amplifier module on it, you will have no trouble removing the amplifier for use in some new project.

Good bypassing of the power supply pin is the sec-

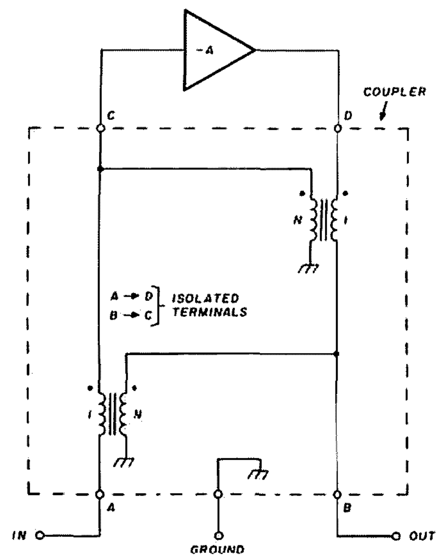


fig. 5. Transformer-type feedback networks provide lower noise figure and higher power to the load than the previous circuits.

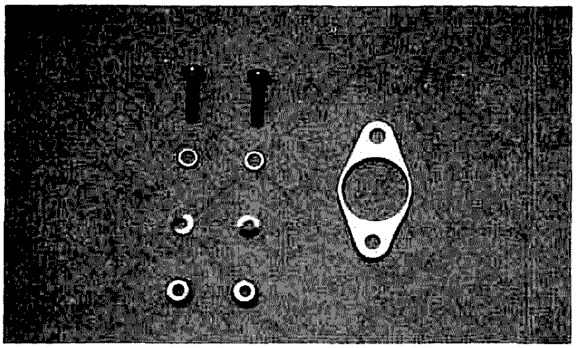


fig. 6. Use of mounting kit aids assembly.

ond thing one must carefully consider in the use of the modular amplifiers. If the pad configuration of **fig. 7** is used, the small chip capacitor (about 100 pF) will provide excellent high-frequency bypassing. However, a larger bypass capacitor (about 10 - 100 μF) is needed to provide low-frequency bypassing. If such a large capacitor is added in parallel with the chip capacitor, the chip may resonate with the lead inductance of the larger capacitor at some frequency. At that resonant frequency, the power supply impedance will vary, causing a glitch in the amplifier response similar to that shown in **fig. 8**. In some cases, this problem can be serious enough to cause oscillation. This problem is easily solved, however, by the addition of a medium-permeability ferrite bead (Ferroxcube 56-590-65/A46 or Amidon FB43101, for example). This bead acts as a lossy element at the resonant frequency. This spoils the Q of the resonant circuit and eliminates the problem. Use of ferrite beads is always a good practice whenever large and small bypass capacitors are paralleled. Be careful, however. Some types of beads are very conductive. These will short the power supply to the groundplane. It is best to use a low conductance bead similar to those above. If you use a highly conductive bead, be sure to insulate it from the lead on which it is placed.

Another point to be considered in bypassing is that of decoupling the supply lead of each module from all others in a cascade. Simple capacitive bypassing of each module with all the V_{cc} leads tied directly to a common supply bus may prove inadequate. A much more effective technique is to feed each module from the supply bus through a small resistance (about 10 ohms for currents up to approximately 100 mA). This resistance aids in decoupling and in providing additional Q spoiling of any resonant networks which may be lurking in long power supply conductors.

amplifiers can be cascaded

Many of the modular amplifiers are designed specifically for cascading. These are unconditionally stable for any combination of input and output VSWR. This eliminates the need for the design of matching networks to assure stability. The specifications are given for a conventional 50-ohm system and the device VSWR is reasonably good. This provides near optimum stage-to-stage matching without the need for matching networks. Several manufacturers offer a series of devices of graduated power capability to allow convenient cascading for both gain and output drive. We will be using two such series of devices manufactured by Avantek in this amplifier design.

A unique feature of these cascadable devices is that the frequency response of the devices is tailored by the manufacturer in such a manner that overall frequency response is preserved with cascading. Normally, if a number of amplifiers with an upper cutoff frequency of f_1, for example, are cascaded, the combined cutoff frequency would be well below f_1. Preservation of a bandwidth is done by a slight peaking of the response near the upper cutoff. The response flatness of a typical cascade will be about ±2 dB. If very flat response is needed (better than ±1 dB), gain may be traded for response flatness by adding compensating attenuator pads between stages as discussed below.

The manufacturers advertise foolproof cascading capability with simple power supply bypassing and direct stage-to-stage coupling. Nevertheless, experience has shown that using these parts in wideband cascaded amplifiers is somewhat more involved than that. This is particularly true for those who do not have the benefit of a machine shop to fabricate intricate enclosures. However, if you carefully follow the suggestions presented here, you should have no trouble.

table 1. Hybrid amplifier specifications.

Guaranteed specifications at 0-50°C (A), −54° to +85°C (B) case temperature; other specifications at 25°C.

Avantek model	frequency response MHz minimum	gain minimum A	gain minimum B	noise figure (dB) typical	reverse isolation (dB) typical	power output for 1 dB gain compression (dBm) typical	3rd order intercept point (dBm) typical	2nd order intercept point (dBm) typical	maximum VSWR (50 ohms) in	maximum VSWR (50 ohms) out	input power volts DC	input power current mA typ
GPD-461	0.1-400	13	12	4.5	20	−2	+9	+9	2.0	2.0	+15	10
GPD-462	0.1-400	13	12	6.0	20	+6	+18	+24	2.0	2.0	+15	24
GPD-463	0.1-400	9	8	7.5	20	+15	+25	+32	2.0	2.0	+24	65
GPD-464	0.1-400	9	8	7.5	20	+15	+26	+38	2.0	2.0	+15	70
GPD-1061	0.1-1000	12	11	6.0	18	+0	+12	+18	2.0	2.0	+15	15
GPD-1062	0.1-1000	12	11	7.0	18	+6	+16	+19	2.0	2.0	+15	27
GPD-1063	0.1-1000	10	9	8.0	18	+14	+24	+36	2.0	2.0	+15	55

Note: Three external capacitors (input, output coupling, and RF bypass) are required to establish low frequency roll-off.

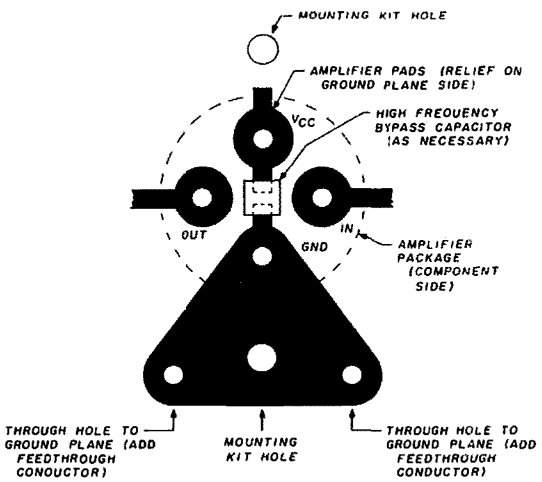


fig. 7. Ground pad configuration details by-pass capacitor, mounting and through hole locations.

design procedure

We are now ready to actually put together an amplifier using the modular hybrid devices. The first step is to choose the actual devices to be used. For the purpose of this project, I have chosen the Avantek GPD 461, 62, 64, and GPD 1061, 62, 63 devices. The actual specifications for these devices are given in **table 1**. The three devices in each series are of graduated power capability and are ideally suited for cascading. Note: there is a GPD 463, but it is a less convenient 24V part. Also, the "60" series parts have no internal coupling capacitors; considerably less costly than their counterparts with internal capacitors, they allow custom tailoring of the low-frequency cutoff. The 400 series parts have a minimum cutoff of 400 MHz and the 1000 series, 1 GHz. This bandwidth is preserved with cascading, and in general it will be found that the upper cutoff is somewhat above that specified. The combined response will be relatively flat, generally ±2 dB, but could be as poor as about ±3 dB. The amplifier design below will allow the purist with access to a network analyzer the means to tailor the bandpass to achieve flatness on the order of ±0.5 dB to beyond 400 MHz, using the GPD 46X parts, and greater than 1 GHz, using the GPD 106X parts.

The manufacturer recommends simple cascading with no matching networks. This technique works reasonably well, but the addition of a simple interstage pad produces a design that is somewhat more forgiving of less than ideal layout and packaging and, in general, is very stable. These pads also tend to reduce the effects of the impedance mismatches from one stage to the next. Furthermore, these pads also provide a very convenient place to add compensation for bandpass flattening. Padding, however, does trade gain for the desirable performance features provided.

I generally use a balanced 50 ohm Pi pad with special conductors on the PC board for "tweaking." A typical schematic (1 dB pad) and layout is shown in **fig. 9**. The additional conductors allow adding chip resistors and capacitors and even small inductors in parallel with the corresponding resistor to provide the frequency response desired. Since gain is traded for performance, it is desirable to keep these pads as small as possible. Typically, 1 dB is adequate if no response tailoring is used. That results in a total 4 dB gain reduction in a three-stage amplifier (two interstage pads, one input pad, and one output pad). That is not too high a price to pay for simplified construction and increased stability.

An input pad also provides some degree of input protection against overdrive. The active device of the modular amplifiers considered here is a bipolar unit. Therefore, it will tolerate a reasonably high current drive in the forward direction. However, in the reverse direction, the emitter-base junction would be avalanched, causing possible damage or degradation to

the part (such as degradation of NF). The emitter-base junction may be protected from avalanche by adding a Schottky diode in parallel with the emitter-base junction, the diode anode at the emitter, and the cathode at the base. In normal operation, this diode is reverse-biased by the base forward drop, about 0.5V. At that bias, the diode impedance is much larger, even at 1 GHz, than the base impedance; it causes little performance degradation. The input pad now provides a well defined minimum source impedance to limit both forward base current and reverse diode current. Without the input pad, the source impedance could be essentially zero (from a capacitive discharge, for example) resulting in very high currents which could degrade performance.

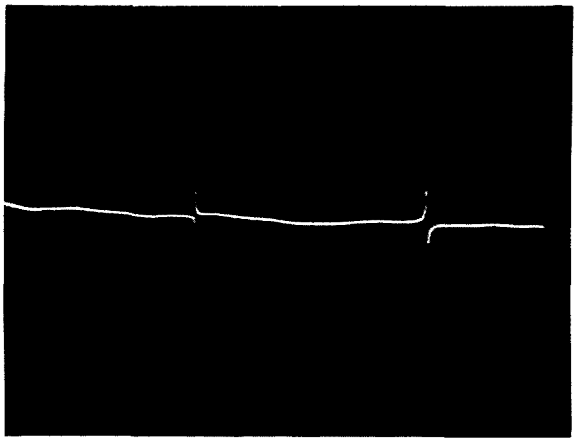


fig. 8. Varying power supply impedance causes glitch in amplifier response.

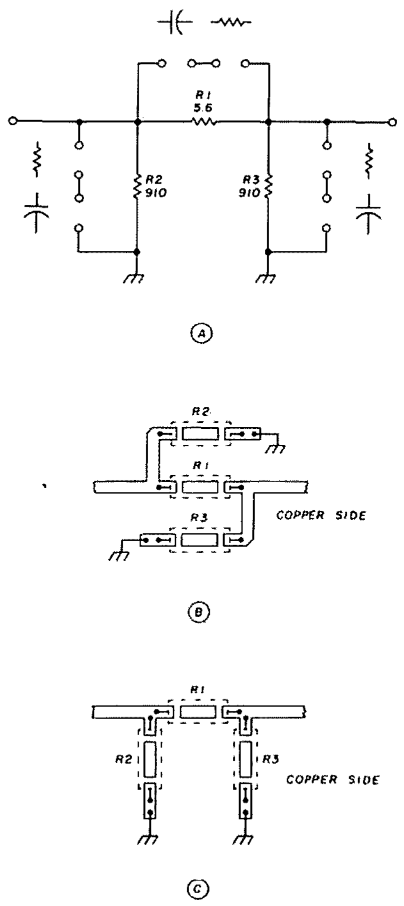


fig. 9. Typical interstage pads: (A) attenuator schematic; (B) input configuration; (C) interstage and output configuration.

The gain and bandpass of the modular hybrids are affected by the operating point. The GPD series parts selected for this project have internal collector load resistors. These basically set the supply potential required for the proper operating point. All the devices chosen are specified for operation at 15V. However, if the supply is varied slightly from 15V, the gain is varied without much degradation in bandpass. The supply potential should not be raised much above 15V due to increased dissipation, but it may be lowered to about 12V, which will reduce the gain of the three-stage amplifier by about 5 dB (**fig. 10**). In this design, a three-terminal regulator is included to allow simple internal regulation of the supply potential for gain stability in the finished amplifier.

packaging

One of the most difficult tasks in homebrew wide-band amplifier design is how to package your product in a way that provides consistent performance. An enclosure machined specifically for the task would be ideal, but because few of us can afford such exotic packaging, I've come to favor the various die-cast aluminum boxes. You're probably familiar with these as being almost right for many uses, but never completely right for anything. Well, this application is no exception. The PC board must be solidly mounted with good groundplane continuity. Simply mounting the PC board on spacers from the bottom of one of these boxes is generally ineffective in RF designs. A better technique is to mount the PC board to a shelf at its periphery, but because this mounting is not provided in die-cast boxes, it must be added. **Fig. 11A** and **B** show a reasonably simple technique for mounting RF boards in these boxes. This technique provides good groundplane coupling to the box and good ground continuity from the RF connectors to the board.

It is very difficult to get perfect alignment of all the mounting components, especially with handmade parts. Therefore, give the mounting holes plenty of clearance to allow for adjustment.

where to obtain parts

Most of the components are standard items and can be obtained from almost any supplier. A few, how-

ever, are somewhat troublesome to find because few distributors stock them. Fortunately, there are alternatives for those parts that prove impossible to get.

Obviously, the most critical special components are the RF amplifiers; these are available from Spirit Electronics* and other Avantek distributors. The chip tantalum capacitors are a particular problem. I used the Matsuo parts, but Matsuo has no retail distributor and no means of handling small orders. The Sprague 193D parts are equally as suitable and are available from some distributors. Another chip capacitor that works quite well is the Alchip-S unit by United Chemi-Con.** However, this part is an aluminum electrolytic. If you use this part, you must *not* clean the PC board in trichlorethylene because the solvent may ruin the capacitor; use 91 percent isopropyl alcohol only. If you cannot find any of those chip capacitors, the dipped tantalum (196D or equivalent) or the CK05 ceramic (with increased low frequency cutoff) can be used with almost as good a result.

The Schottky diode for input protection can be a problem. Almost any small signal RF Schottky diode is suitable. The Radio Shack 276-1124 is a good substitute. If a suitable diode cannot be found, just omit it, but this will make the input device more sensitive to reverse bias damage.

The swagged terminals are quite convenient and give the board a professional appearance, but are usually not available in small quantities. Proper swagging of these is essential; if they are too difficult to obtain or proper swaggging cannot be done, use pieces of wire as suggested below.

components

Because this is a high frequency design, performance is very sensitive to the particular components used. The components specified in the list of materials are the specific ones used for construction of the units presented in this article. While other components can be used with equal success, care must be exercised in their selection. Several points to take into consideration in selecting alternative parts are given below.

The components most easily substituted are the resistors. I specified Allen-Bradley brand carbon composition resistors mainly because of my particularly good experience with these parts. Most other name brands should perform equally well. However, some carbon film components are spirally cut to obtain the desired resistance; this could introduce added inductance, which could perturb performance. Also, some of the less expensive parts exhibit excess noise that could increase the noise figure of the amplifier. In general, try to avoid using the very inexpensive resistors and use brand name parts instead. In the 400

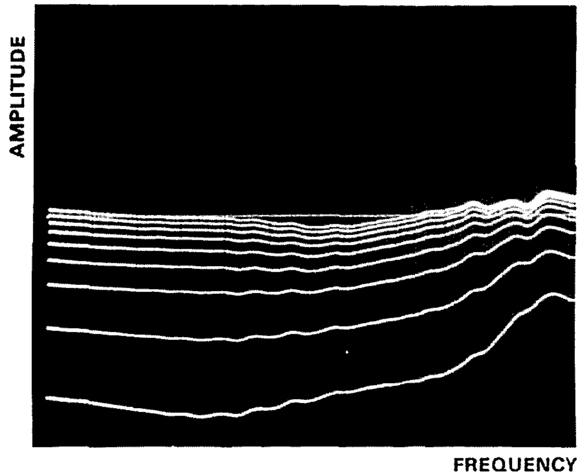


fig. 10. GPD106X gain variation with supply voltage (without regulator). Gain changes in 1 dB steps as V_{CC} is varied from +8 to +16 VDC.

MHz amplifier (GPD 46X amplifiers), either carbon composition or quality carbon film resistors should be suitable, but in the 1000 MHz unit (GPD 106X amplifiers), try to use only carbon composition resistors.

The interstage coupling capacitors are also good candidates for substitution because the ones specified are somewhat difficult to find. If low frequency response (i.e., below about 100 kHz) is not a consideration, standard CK05 capacitors work quite well. These are not normally recognized as being good high-frequency capacitors, but up to 1 GHz or so they will work very well. (Fig. 12 shows typical performance of the 1 GHz amplifier with 0.1 μF CK05 coupling capacitors.) One component almost guaranteed *not* to work for coupling is the tubular aluminum electrolytic capacitor, which generally exhibits high lead inductance and high stray capacitance because of its size; its use will probably result in a very unstable amplifier. The chip aluminum electrolytic capacitor manufactured by United Chemi-Con seems to work well, but is equally as difficult to obtain as the chip tantalum capacitors specified. With a little care, the dipped tantalum units — i.e., Sprague 196D or equivalent — can be used successfully for coupling. The general key to success with these parts is to stick to the ones of small physical size, about 1/8 inch in diameter, and keep leads short. An operating voltage greater than 6.8V is necessary. Fig. 13 shows the response of the 1 GHz amplifier with 10 μF, 20V 196D capacitors.

The other capacitors in the circuit are not too critical, but some care should be taken in their selection. The low-value bypass capacitors should have relatively good high frequency properties. The CK05 units are recommended. Depending on their manufacturer, typical ceramic discs vary tremendously in performance and are not recommended. (The actual value

*Spirit Electronics, 6560 N. Scottsdale Road, Suite E204, Scottsdale, Arizona 85253

**United Chemi-Con, 9801 West Higgins Road, Rosemont, Illinois 60018

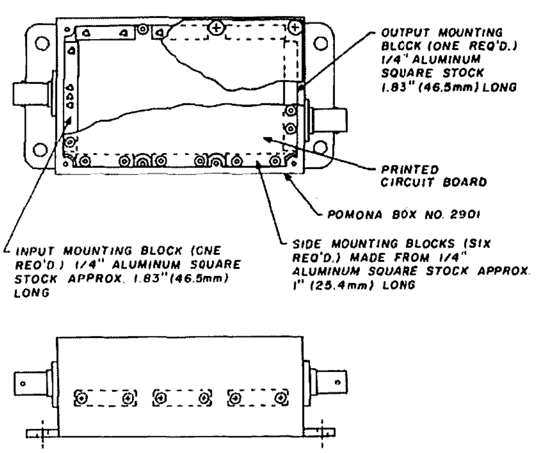


fig. 11A. Mounting shelves provide good ground plane coupling to box and good ground plane continuity between RF connectors and board.

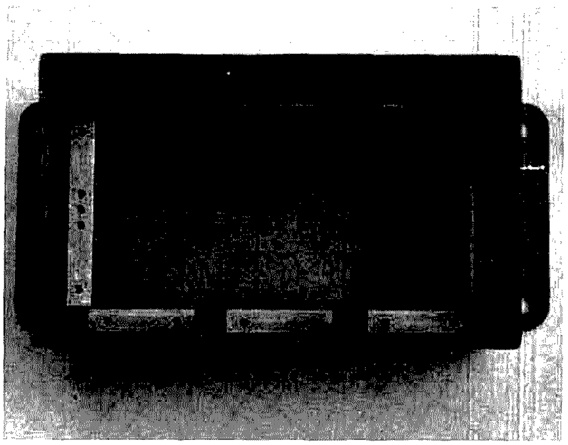


fig. 11B. Box detail with mounting blocks installed.

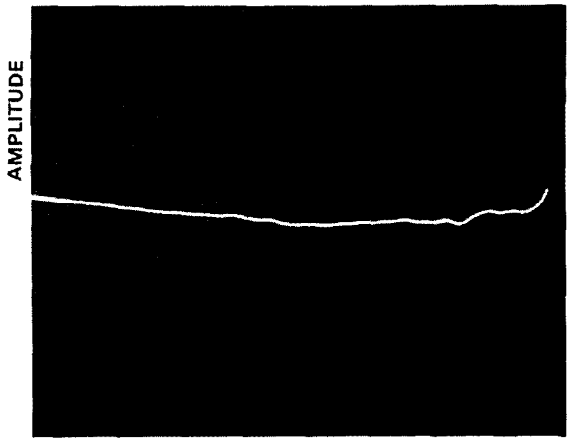


fig. 12. Typical performance of 1 GHz (0 - 1200 MHz, 5 dB gain variation) amplifier using 0.1 μF CK05 coupling capacitors.

is not too critical since they are basically used for high-frequency bypassing. Values between 100 and 1000 pF should be good.) The higher value bypass capacitors for both the amplifiers and the regulator should be dipped tantalum. Don't forget to bead the positive lead of the three amplifier bypass capacitors or you may have response anomalies similar to those shown in **fig. 8**. Aluminum electrolytics are too large to fit conveniently in the enclosure. Notice on the PC board that there are two ground pads for each of these capacitor positions; this allows easy mounting of both 0.2 and 0.25-inch lead formed capacitors. The actual value of these large bypass capacitors is not too critical. Variations by ±20 percent should not be a problem, but it is better to go higher than lower. Also, observe the operating voltage. All the units except the supply input bypass capacitor should be 16V or greater. Try to stick with 16V capacitors since higher voltage units are larger and more expensive. The input capacitor should be a 35V unit to allow for large supply voltages without damage.

Another area that can prove troublesome is the swagged terminals. If you use terminals, they must be properly swagged and soldered into the board. If they are not properly swagged, the solder connection on the trace side could become a cold solder joint when the terminal is heated to attach leads. If you cannot properly swage the terminals, it is better to use pieces of wire in their place. A ¼-watt resistor lead works well for two RF connections (keep them as short as possible), and a lead from the power supply reverse polarity protection diode, CR2, works well for the power supply terminal. No ground terminal is really needed on the board; it's included only for checkout convenience. If you want to use terminals, almost any type will be adequate, but the PC board holes may have to be changed.

The ground terminal on the outside of the box may prove difficult to find. If you cannot find these, use a ¾-inch 4-40 brass or plated steel screw with a nut. Place the nut on the screw about ¼-inch from the end. Drive the screw into the box ground lug hole about ¼-inch and lock it in place with the nut. This will allow the screw to extend out about ½-inch for lead attachment.

Finally, the feedthrough filter could be a problem. While almost any filter will work, some type of feedthrough filter or feedthrough capacitor should be used; it must be a threaded type because it's difficult to solder to the box alloy. If you use a filter different from the one specified, be sure to drill and tap the mounting hole for the specific filter selected rather than for the filter shown in the list of materials.

PC board assembly

The assembly of the PC board is reasonably straightforward. (The schematic, component and PC

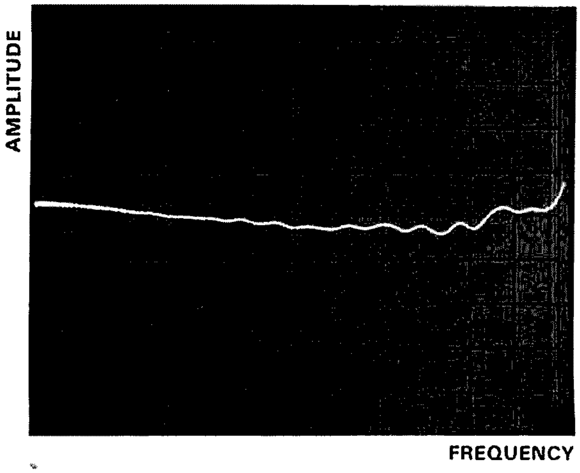


fig. 13. Response of 1 GHz (0-1200 MHz, 5 dB gain variation) amplifier using 10 μF, 196D capacitors.

boards are shown in **figs. 14, 15,** and **16.**) The first thing to do is mount the swagged terminals if you choose to use them. Next mount the three amplifiers with the mounting kits provided. Be sure to have the mounting kits completely installed and tight before soldering any of the amplifier leads. If chip tantalum coupling capacitors are to be used, it is easiest to mount those next. (These mount on the trace side of the PC board.) Now mount the remainder of the components on the groundplane side as shown on the assembly drawing, paying particular attention to the groundplane solder connections. Be sure to apply enough heat to the groundplane to get good solder flow on it.

When mounting the voltage regulator, allow about 1/16-inch space between the regulator body and groundplane. Mounting this part may be somewhat confusing because there are two sets of pads. The set closest to the board edge is for the general 78XXT and LM340T series parts. The other set is for the LM317T. (This allows the use of the LM317, 7805, 7812, 7815 or equivalent parts — the LM317 is preferred.) To mount the regulator, the leads must be bent as shown in **fig. 17**, which shows the lead bending for the LM317. If a 78XXT or LM340T series part is used, the leads should be bent behind the body to keep the mounting position of the tab at the proper location. Be careful to keep the tab electrically isolated from both the groundplane and the box. The two programming resistors in the regulator circuit are shown for the LM 317. If a 78XXT part is used, these resistors must be replaced with the appropriate values from the table on the schematic. These, however, are only approximate values. The supply potential at the amplifiers should be checked and the resistor values adjusted, if necessary, to set the output amplifier supply voltage between 15V and 16V. Usually only resistor R16 needs to be adjusted slightly. Increasing R16 lowers the output voltage, and decreasing R16 raises it.

fabricating the enclosure

Building the enclosure is probably the most difficult part of this project. Because the final performance will be as much a function of the enclosure as the actual PC board, considerable care should be invested in this task.

First the paint must be removed from the interior of the box to allow good grounding of the PC board. (The easiest way to obtain effective grounding is to buy the box unfinished, but this results in a finished unit that is less attractive than it could be.) If you start with a finished box, fill it with paint remover to within about one-half inch of the top. (Be sure to use proper

VR1	R10	R17
LM317 (PREFERRED)	330	3.9k
LM340T-5	470	620
LM340T-12	1k	200
LM340T-15	1k	24
LM7805	470	560
LM7812	1k	180
LM7815	1k	22

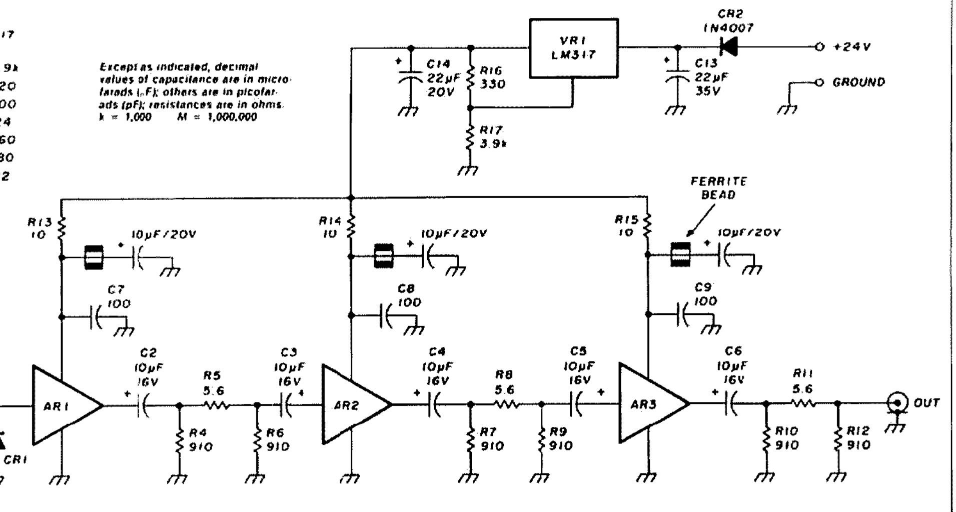


fig. 14. RF amplifier schematic.

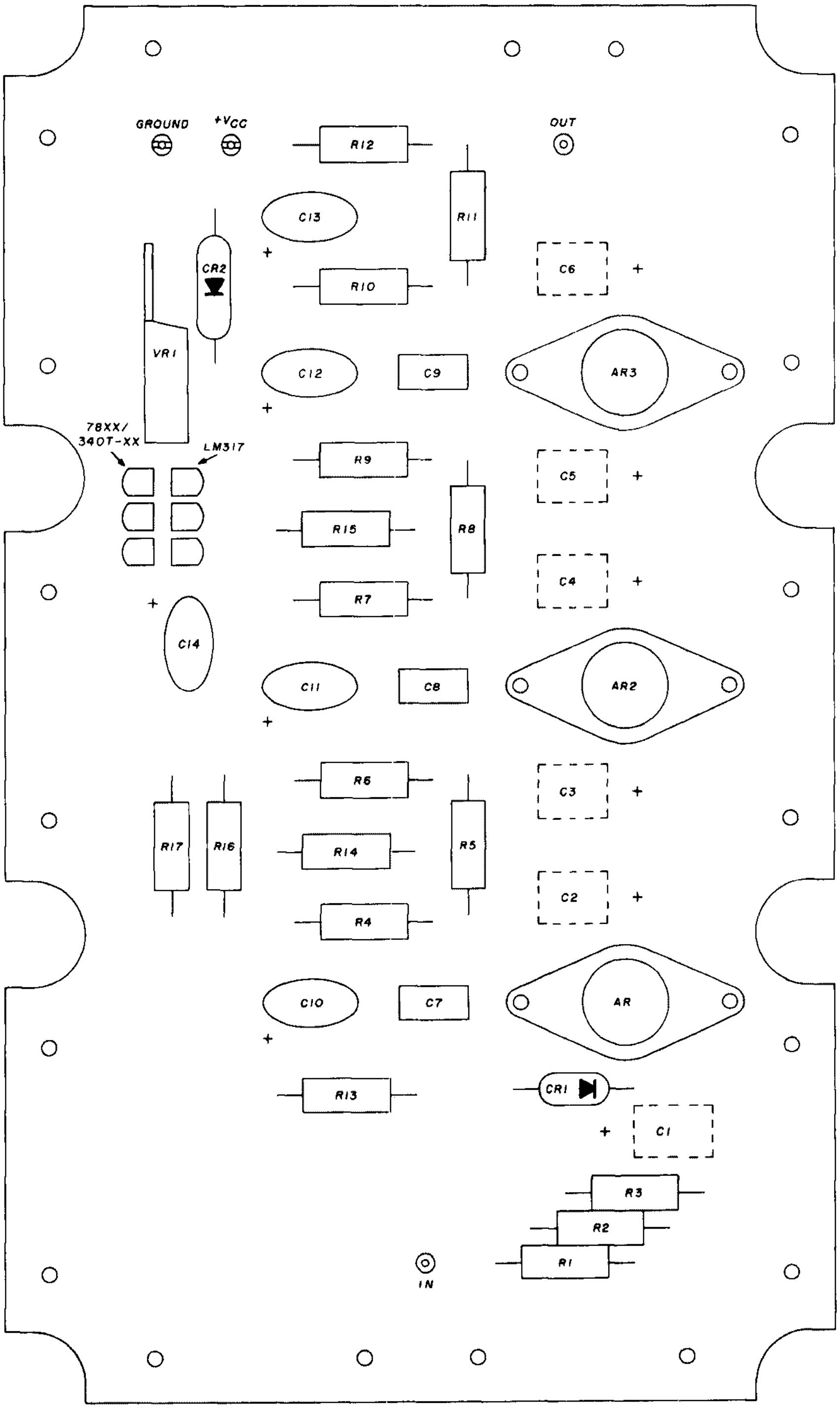


fig. 15. PC board component layout.

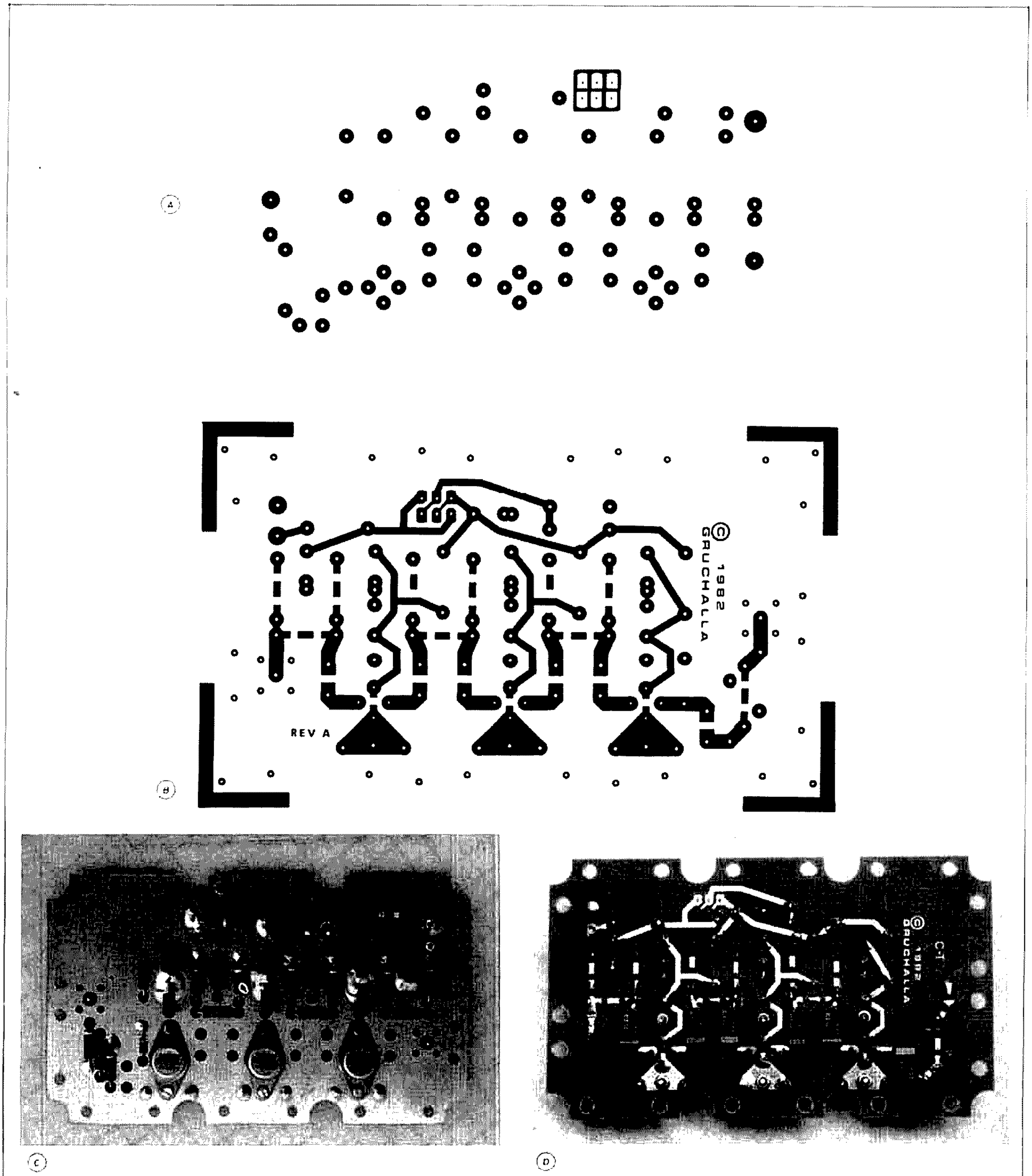


fig. 16. Double-sided PC board: *(A)* component side artwork, *(B)* ground plane side artwork, *(C)* completed assembly — component side view, and *(D)* completed assembly — ground plane side.

eye protection and gloves when handling paint remover.) Leave the paint remover in the box until it begins to attack the paint. Next, pour the paint remover out and *immediately* rinse the box in fresh paint thinner. Follow this with a thorough scrubbing with soap and water; a hard spray from a garden hose will help remove the loosened paint. Repeat the process until the paint has been removed. (If you prefer, you can remove the paint by sanding the interior walls of the box.)

Materials list.

item	description
AR1	Avantek GPD 1061 (or GPD 461)
AR2	Avantek GPD 1062 (or GPD 462)
AR3	Avantek GPD 1063 (or GPD 464)
ferrite bead	Arnidon FB 64-101 or Ferroxcube 56-590-65/4AG
C1-C6	10 μF/6.8V chip tantalum capacitor (may substitute CK05 ceramic or 196D dipped tantalum)
C7-C9	100 pF - 1000 pF, CK05 ceramic capacitor
C10-C12	10 μF/20V, dipped tantalum capacitor
C13	22 μF/35V, dipped tantalum capacitor
C14	22 μF/20V, dipped tantalum capacitor
CR1	1N5711 Schottky diode (may substitute Radio Shack 276-1124)
CR2	1N4001 Diode (may substitute 1N4002-1N4007)
filter	Erie 1221-001
R1,R3,R4,R6 R7,R9,R10,R12	910 ohm, 1/4 watt, 5 percent, RC07 resistor (Allen Bradley preferred)
R2,R5,R8,R11	5.6 ohm, 1/4 watt, 5 percent, RC07 resistor (Allen Bradley preferred)
R13-R15	10 ohm, 1/4 watt, 5 percent, RC07 resistor (Allen Bradley preferred)
R16	330 ohm, 1/4 watt, 5 percent, RC07 resistor (LM317 only)
R17	3.9K ohm, 1/4 watt, 5 percent, RC07 resistor (LM317 only)
VR1	LM317 regulator (may substitute as *shown on schematic* — R16 and R17 must also be changed)
box	Pomona 2901 (may substitute 2906)
terminal	(2) Cambion 160-2081-02-01-00 (input and output)
terminal	(2) Cambion 140-1385-02-01-00 (power and ground)
terminal	(1) H.H. Smith 2009 (box ground)
RF connector	UG447/U (may substitute UG290A/U with mounting holes drilled out to 7/64)
hardware for mounting regulator	
spacer	H.H. Smith 2341, 3/8 × No. 4 brass
insulator	(1) mica
shoulder washer	(1) fiber
screw	(1) 4-40 × 5/8, pan head steel
other hardware	
screw	(34) 2 × 56 × 3/16, pan head, steel (PC board and box)
screw	(8) 2 × 56 × 3/8, pan head steel (connector)
flat washer	(42) No. 2
flat washer	(1) No. 4
aluminum bar	12" 1/4 × 1/4

The various holes must be added next. Use the PC board hole pattern (**fig. 16**) as a guide to hole placement in the mounting stock. For those who wish to precisely replicate hole locations, a detailed mechanical drawing is available from *ham radio* (enclose SASE with request). If you lay the holes out by hand, it is best to use a precision rule with both fraction and decimal graduations. Carefully mark all hole positions with a sharp scribe. Then check the dimensions to make sure all holes are in their correct locations. Be sure to center-punch all holes before drilling to prevent the drill from "walking." After drilling, deburr all the holes.

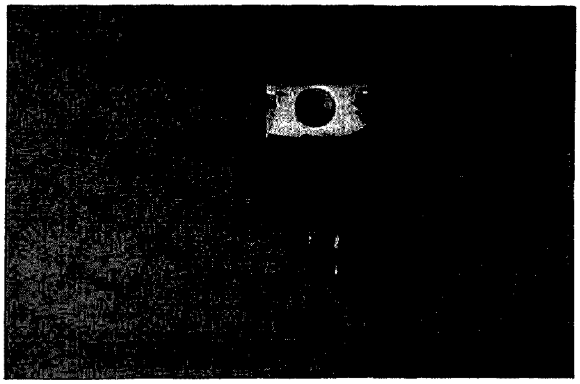


fig. 17. Regulator leads are bent for ease of mounting.

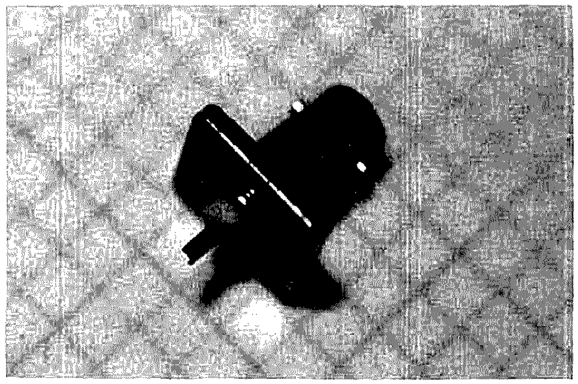
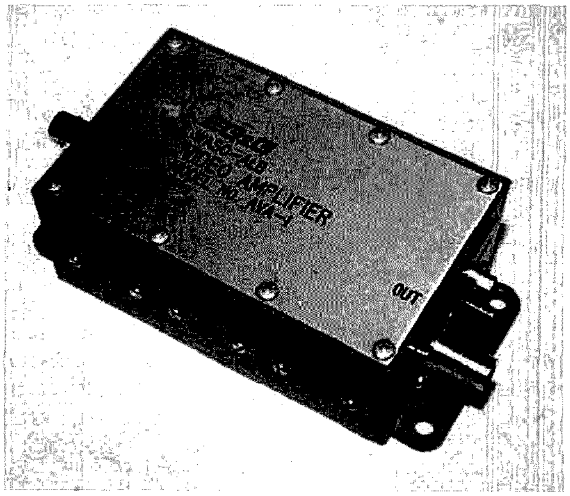


fig. 18. Dielectric ring of RF connector is modified to fit.

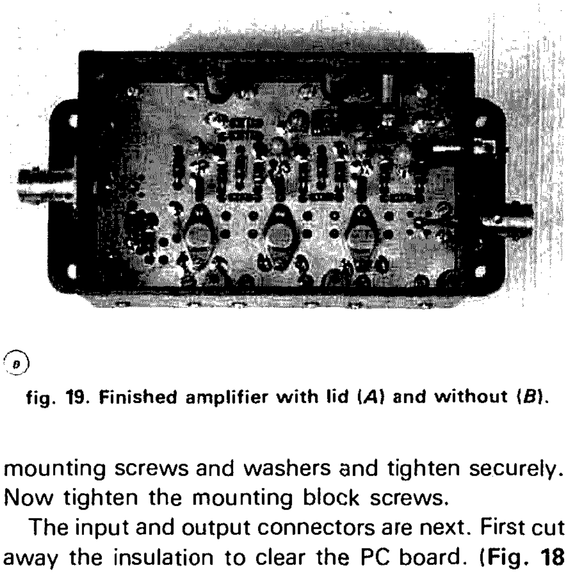
Now the mounting blocks must be made. The choice of material is not too critical. The 1/4-inch square material is specified, but 1/4 × 3/8 or 1/4 × 1/2 could also be used. Suitable mounting material — and the taps as well — should be available from your local hardware store. Carefully mark and cut each of the eight mounting blocks. Then mark, center-punch and drill each of the necessary holes. (Use motor oil or some other light oil to lubricate the bit.) Then tap each hole. Be *very* careful tapping. Use oil generously and back the tap out often to prevent jamming. This is particularly critical on the 2-56 holes.

final assembly

Now that all the pieces are finished, only final assembly remains. First, mount the eight mounting blocks with screws and washers inside the box, being sure to position each of the end pieces on the correct ends. Leave all screws just finger-tight to allow some adjustment. Place the PC board in the box on the mounting blocks. Again, make sure the orientation is absolutely correct. Insert all the PC board



(A)



(B)

fig. 19. Finished amplifier with lid (*A*) and without (*B*).

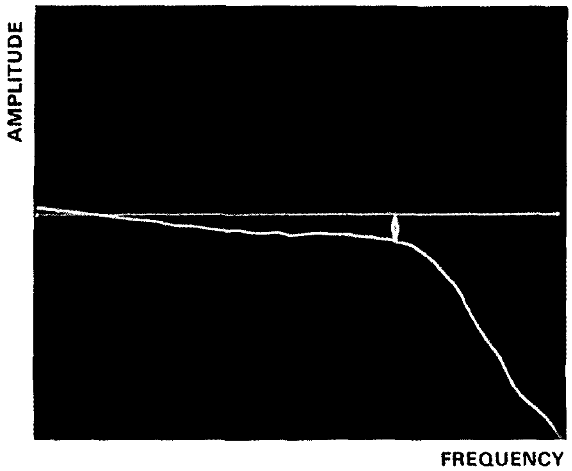


fig. 20. Swept frequency response of HF amplifier using GPD 460 series devices; the marker "birdie" is at 825 MHz (− 3 dB point).

mounting screws and washers and tighten securely. Now tighten the mounting block screws.

The input and output connectors are next. First cut away the insulation to clear the PC board. (**Fig. 18** shows this modification.) Slide a knife along the solder tail into the insulation to the flange of the connector. Next, cut parallel with the flange to the solder tail and remove the piece of insulation. The connectors may now be mounted. Attach the conductors from the solder tails to the corresponding terminals. Then mount the ground terminal and power filter. Wire the power filter to the power terminal. Mount the lid and you're done.

performance

When you're finished, your amplifier should look like the one pictured in **fig. 19**. While having a machine shop available helps tremendously, the unit shown — except for the engraved lid — was build by hand exactly as described above.

Amplifiers were built using both the GPD 461/462/464 parts and the GPD 1061/1062/1063 parts. The first of these is the lower frequency unit. The bandpass of this amplifier is shown in **fig. 20**. Although the GPD 46X series parts are specified as 400 MHz devices, **fig. 20** shows that they perform far better. This unit was flat within ±2 dB from **500 Hz** to almost 900 MHz. The nominal gain of this unit was 34 dB. If you have a suitable network analyzer available, some of that gain could be traded off to allow compensation of the gain to flatten bandpass. *The maximum output level of this amplifier at the* 1 dB compression point was +14 dBm. The noise figure (NF) was measured roughly as 2.8 dB (260 degrees K). This is considerably better than that specified for the GPD 461 first stage, but this is not unusual because the specification value must be conservative enough to allow the manufacturer a reasonable yield. The overall dynamic range with a bandpass of 400 MHz, 2.8 dB noise figure, and +14 dBm 1 dB compression point, is then greater than 97 dB. This is reasonably high performance in any terms, but particularly good for a homemade unit.

If you think *that* was good, look at the performance of the high frequency amplifier that uses the GPD 1061/1062/1063 devices. **Fig. 21** shows the frequency response of this amplifier. This unit is flat within ±2 dB from 500 Hz to in excess of 1200 MHz, with a gain of +34 dB. The ripple above about 1000 MHz in **fig. 21** is actually due to the test system as shown in the measurement system response illustrated in **fig. 22**. The output at 1 dB compression was +13 dBm. The NF was measured to be about 3.3 dB (330 degrees K). The dynamic range is then greater than 91 dB.

Because these noise figure values seem quite good, one might suspect error in measurement; then too, the method used to measure the NF wasn't the best, but was convenient. To check the accuracy of the measurement system, the NF of an AWL-1200 com-

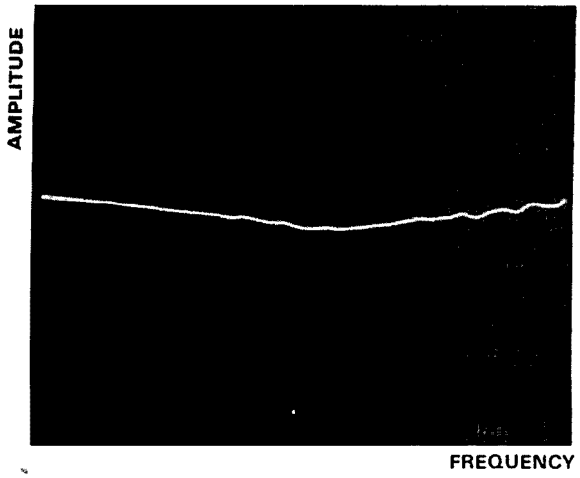


fig. 21. Swept frequency response of HF amplifier using GPD 1060 series devices.

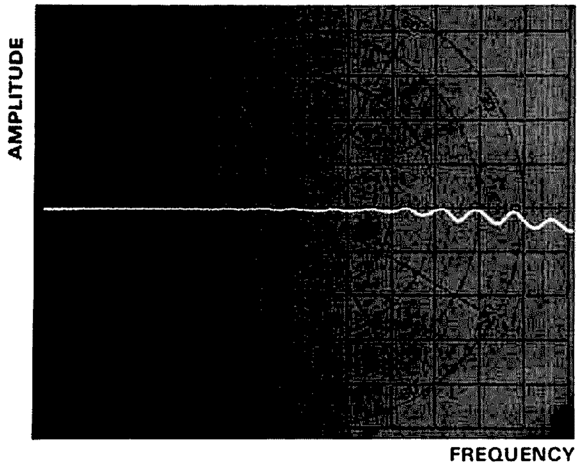


fig. 22. Test instrumentation actually introduces major ripple component.

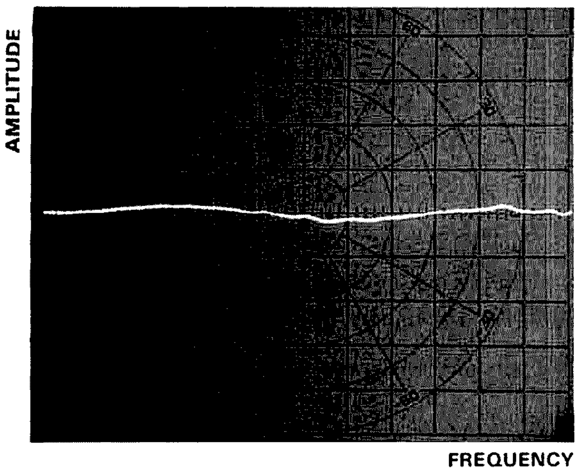


fig. 23. A wideband amplitude "flatness" of ±1 dB is achieved through the use of bandpass compensation.

mercial amplifier was measured. This unit was specified to have a NF of 5 dB and a value of 4.3 dB, was measured. This shows quite good agreement and that the results are reasonably accurate.

To demonstrate bandpass compensation, the high frequency unit was compensated to provide a flatness of ±1 dB. This is shown in fig. 23. About 4 dB of gain was traded for this performance. With enough patience, the bandpass could be flattened even further, but ±0.5 dB or so is about a practical limit.

Fig. 10 shows the gain variation of the high-frequency amplifier with operating voltage. For this example, the regulator output voltage was varied from 16V down to 8V in 1V steps with the highest gain corresponding to the highest voltage. This shows that some slight gain control can be provided with variation of supply voltage.

Performance of the high-frequency amplifier with CK05 0.1 μF coupling capacitors is shown in fig. 12. This raised the low frequency cutoff to about 50 kHz. Notice the slight peak in response at 1200 MHz in fig. 12. This could be attributed to stray coupling or the poorer impedance characteristics of the CK05. This peak was found to reach a +5 dB maximum at 1290 MHz beyond which the response dropped normally. Fig. 13 shows the response with 196D, 10 μF, 20V dipped tantalum coupling capacitors. The low frequency cutoff was again 500 Hz and the high frequency characteristics were much the same as the CK05 — not perfect, but certainly acceptable.

conclusions

While constructing the amplifier described is not easy, it certainly should be within the capability of most readers. Once past the mechanical construction, the electronic assembly is simple. (This is quite the opposite of most RF amplifier construction projects.) Exercising care and patience in building this unit results in excellent performance; furthermore, the cost is quite modest compared to commercially available units offering similar performance.

I believe this project will demonstrate the simplicity and convenience of designing with modular hybrid amplifiers. This will be particularly obvious to those adventurous individuals who've spent many hours with plastic tweezers and razor blades, "tweaking" a discrete design. (Even when finished with those labor-intensive designs, you could not be sure of performance without a network analyzer.) If you build this amplifier carefully, its performance will almost certainly be similar to that documented above. And, if you follow the design hints given above, you should have little trouble using modular hybrids in your own designs.

reference

1. C.D. Motchenbacher and F.C. Fitcher, "Low Noise Electronic Design," *John Wiley & Sons*, New York, N. Y. 1973.

ham radio

static electricity
and
modern integrated circuits

To prevent damage or destruction, ground *yourself*

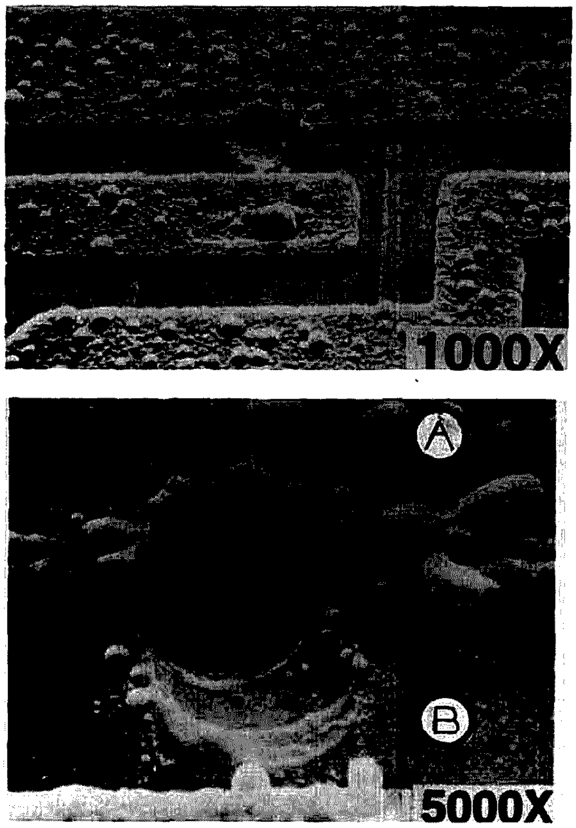


fig. 1. Enlarged, detailed views of a 6-micron (0.0002 inch) diameter hole created in aluminum metallization (A) and silicon dioxide substrate (B) by static electricity on an op amp integrated circuit.

By Morris H. Lundberg, K4KEF, 131 Burnett Way, Alpharetta, Georgia 30201

The integrated circuit is no exception to the familiar rule that says you can't get something for nothing. While the use of ICs has enabled manufacturers to reduce both size and power requirements of battery-operated radio equipment, there is a price for this increased portability and convenience. That price is extreme vulnerability to static electricity (**fig. 1**).

But are ICs really as sensitive to static damage as suggested in product packaging and literature?

The answer is *yes*.

A person can develop static levels of several thousand volts merely by walking across the floor. A discharge of this intensity can destroy or seriously degrade sensitive electronic components or circuit boards. Degraded boards may be prone to premature failure.

During routine maintenance and troubleshooting, the General Telephone Company of Wisconsin found evidence of degradation and premature failure of electronic telecommunication equipment due to static discharge in handling.[1] It also found that circuits were being damaged during handling for storage. To eliminate the problem, General Telephone initiated an Electrostatic Discharge (ESD) program.

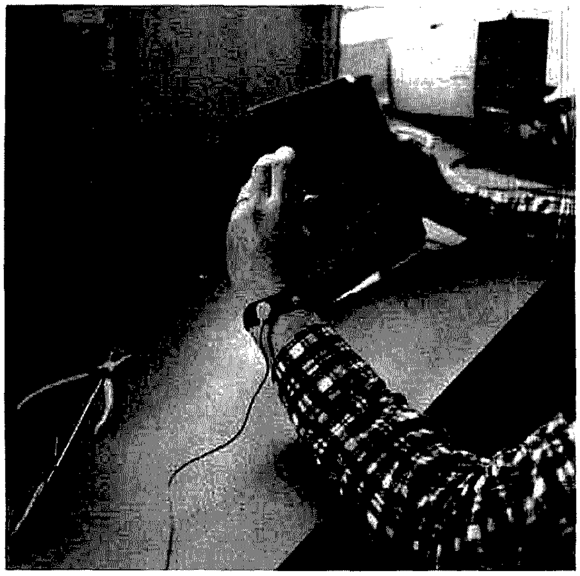


fig. 2. Conductive wrist strap prevents damage to sensitive electronic circuitry by draining static buildup from worker. Humans can build up more than 8000 volts of static charge on themselves with normal activity. Some sensitive devices, however, can be destroyed or degraded by less than 100 volts of static charge.

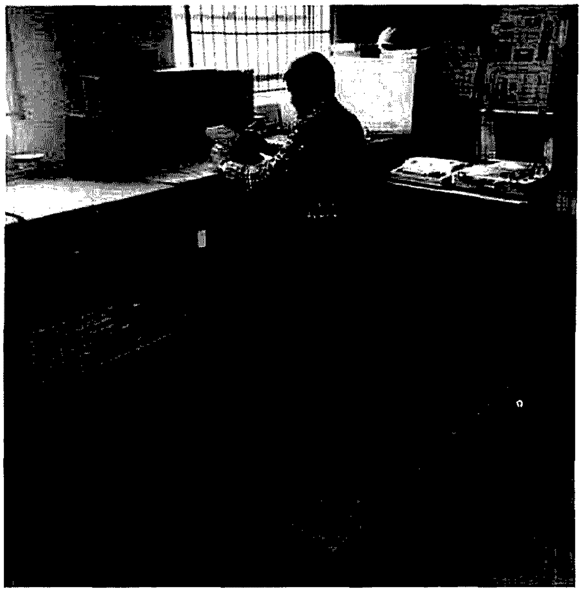


fig. 3. Work stations, where delicate electronic parts and circuits are handled, should be protected against static electricity. Protection includes conductive table and floor mats, as well as wrist strap pictured in *fig. 2*. A blower also circulates a stream of negatively charged ionized air to neutralize static on non-conductive materials, such as plastic coffee cups, which cannot be grounded.

Semiconductor manufacturer Mostek also developed an ESD program in which it was able to reduce catastrophic ESD failures in the devices most susceptible to static damage by about two thirds on final test line production. In one Hewlett-Packard packaging plant, the yield went from 25 to 100 percent, with an overall improvement of 10 percent in IC lab yields after ESD procedures were instituted.[2] Tel-Matic Systems of Toledo, Ohio, reported a 60 percent reduction in the failure rate of newly installed electronic telephone systems after an ESD program was initiated.[3]

Although elaborate and costly, the ESD programs conducted by these companies consist primarily of static grounding of all personnel who handle sensitive devices, as well as the institution of improved handling, storage, and shipping methods. Several techniques of static grounding are used. Grounded wrist straps (**fig. 2**) and grounded, conducting workbench and floor mats (**fig. 3**) drain away static charge before any damaging voltage can develop.

Anti-static plastic pouches have been used for some

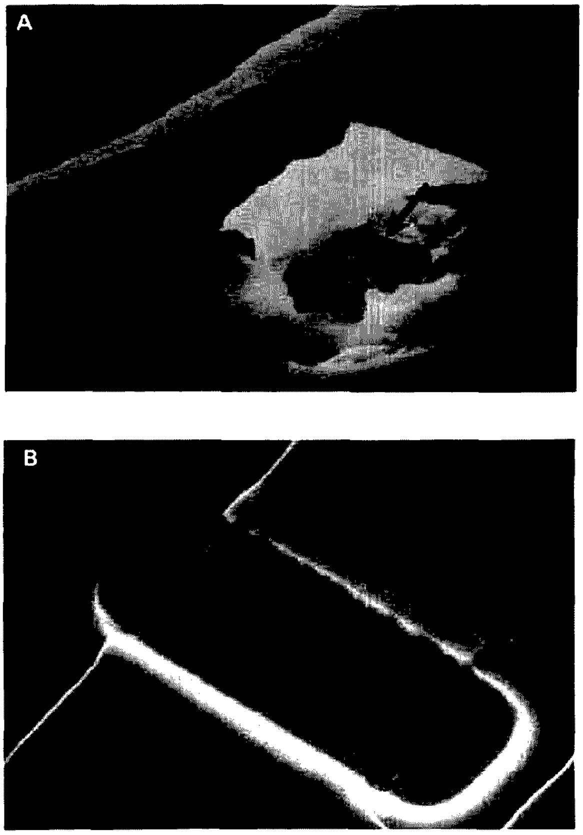


fig. 4. (A) shows static damage, magnified 7000 times, to a 3N157 MOSFET; (B) shows damage to an input pull-up resistor on a p-MOS character generator at 3000x magnification.

moon-tracking by computer

Determine azimuth and elevation with simplified computer program

A simple, user-friendly program of 58 steps to determine the position of the moon is now available for the TRS-80C* computer. Based on a more complex moon tracking program developed by the EIMAC division of Varian, Inc., this program computes and prints out the azimuth and elevation of the moon for every quarter-hour.

There's no list of rules to be followed and the entries are very short. Only six are required:

- LATITUDE OF LOCAL STATION
  (degrees, minutes, North or South)
- LONGITUDE OF LOCAL STATION
  (degrees, minutes, East or West)
- GHA-1 AT EVEN HOUR BEFORE MOONRISE†
  (degrees, minutes)
- GHA-2 12 HOURS LATER
  (degrees, minutes)
- DEC-1 AT EVEN HOUR BEFORE MOONRISE‡
  (degrees, minutes, North or South)
- DEC-2 12 HOURS LATER
  (degrees, minutes, North or South)

Each function is assigned a letter; the long mathematical equations then become simple expressions of letters (see **table 1**). This reduces the chance of error considerably. The computations then require only about a dozen steps, using a straight-line interpolation of the GHA and DEC values over the twelve-hour period. The AZ-EL accuracy obtained by this simplification is adequate.

table 1. Letter assignments of variables.

A = LATITUDE OF LOCAL STATION
B = LONGITUDE OF LOCAL STATION
C = GHA-1
D = GHA-2
E = DEC-1
F = DEC-2
G = GMT OF MOONRISE
H = (D-C)/48 INTERPOLATION INCREMENT GHA (0.25 HOUR)
I = (E-F)/48 INTERPOLATION INCREMENT DEC (0.25 HOUR)
J = B-C
K = COS(J/Z)
L = COS(A/Z)
M = COS(E/Z)
N = SIN(A/Z)
O = SIN(E/Z)
P = TAN(A/Z)
Q = ARCSIN(X) ELEVATION ANGLE
R = COS(Q/Z)
S = TAN(Q/Z)
T = ARCCOS(Y) AZIMUTH ANGLE
X = (K*L*M) + (N*O)
Y = (O/(R*L)) – (P*S)
Z = 57.2957795 RADIAN CONVERSION FACTOR.

THE PROGRAM IS ITERATED IN QUARTER HOUR INCREMENTS AND MAY BE STOPPED AT ANY DESIRED POINT.

*TRS-80C is a trademark of Tandy, Inc.
†Greenwich Hour Angle, equivalent to longitude on the Earth, is the angular distance of a celestial body west of the celestial meridian of Greenwich.
**Declination, equivalent to latitude on Earth, is the angular distance of celestial body north or south of the celestial equator. (Extension of plane of Earth's equator.)

By I.L. McNally, K6WX, 26119 Fairlane Drive, Sun City, California 92381

TRS-80C Moon-tracking program (LOAD "EME/BAS:0")

```
5 PRINT#-2,"PROGRAM FOR MOON TRACKING. INPUTS ARE IN DEGREES AND MINUTES."
10 INPUT "LATITUDE:DG,MN,N OR S";AA,AB,AC$
15 PRINT#-2, AA;AB;AC$,"LATITUDE IN DEGREES AND MINUTES."
20 IF AC$="S" THEN 25 ELSE 30
25 A=(AA+AB/60)*(-1):GOTO 35
30 A=AA+AB/60
35 PRINT#-2, A "LATITUDE IN DECIMAL DEGREES.(A)"
36 INPUT "LONGITUDE:DG,MN,E OR W";AD,AE,AF$
37 PRINT#-2, AD;AE;AF$, "LONGITUDE IN DEGREES AND MINUTES."
40 IF AF$="E" THEN 45 ELSE 50
45 B=(AD+AE/60)*(-1):GOTO 55
50 B=AD+AE/60
55 PRINT#-2, B "LONGITUDE IN DECIMAL DEGREES.(B)"
60 INPUT "GHA-1 AT EVEN GMT BEFORE MOONRISE:DG,MN";AG,AH
65 PRINT#-2,AG;AH, "GHA-1 IN DEGREES AND MINUTES."
70 C=AG+AH/60
75 PRINT#-2, C "GHA-1 IN DECIMAL DEGREES.(C)"
80 INPUT "GHA-2 12 HOURS AFTER MOONRISE:DG,MN";AI,AJ
85 PRINT#-2,AI;AJ,"GHA-2 IN DEGREES AND MINUTES."
90 D=AI+AJ/60
95 PRINT#-2,D "GHA-2 IN DECIMAL DEGREES.(D)"
100 INPUT "DEC-1 AT EVEN GMT BEFORE MOONRISE:DG,MN,N OR S";AK,AL,AM$
101 PRINT#-2,AK;AL;AM$,"DEC-1 IN DEGREES AND MINUTES."
102 IF AM$="S" THEN 105 ELSE 110
105 E=(AK+AL/60)*(-1):GOTO 115
110 E=AK+AL/60
115 PRINT#-2, E "DEC-1 IN DECIMAL DEGREES.(E)"
120 INPUT "DEC-2 12 HOURS AFTER MOONRISE:DG,MN,N OR S";AN,AO,AP$
125 PRINT#-2, AN;AO;AP$, "DEC-2 IN DEGREES AND MINUTES."
130 IF AP$="S" THEN 135 ELSE 140
135 F=(AN+AO/60)*(-1): GOTO 145
140 F=AN+AO/60
145 PRINT#-2,F "DEC-2 IN DECIMAL DEGREES.(F)"
150 INPUT "STARTING GMT"; G
160 Z=57.2957795
165 H=(D-C)/48
175 I=(E-F)/48
182 PRINT#-2, G "GMT"
185 J=B-C
190 K=COS(J/Z)
195 L=COS(A/Z)
200 M=COS(E/Z)
210 N=SIN(A/Z)
215 O=SIN(E/Z)
220 P=TAN(A/Z)
225 X=(K*L*M)+(N*O)
230 Q=Z*(ATN(X/SQR(-X*X+1)))
231 PRINT#-2,Q "ELEVATION ANGLE (Q)"
235 R=COS(Q/Z)
240 S=TAN(Q/Z)
245 Y=(O/(R*L))-(P*S)
250 T=Z*(-ATN(Y/SQR(-Y*Y+1))+1.5708)
255 PRINT#-2, T "AZIMUTH ANGLE.(T)"
260 U=360-T
265 PRINT#-2,U "AZIMUTH ANGLE WHEN GHA>LONGITUDE.(U)"
270 G=G+.25
275 E=E-I
280 C=C+H
285 GOTO 182
```

ham radio

VHF/UHF WORLD

Joe Reisert W1JR

VHF/UHF receivers

It's been some time since my last article on VHF/UHF receivers.[1] In many ways, the substance of that article still stands as written; however, technology marches on, and the state-of-the-art *has* improved: the development of GaAs FETs (gallium arsenide field effect transistors) and the proliferation of commercial transceivers for the VHF and lower UHF regions are just two examples of important changes that have taken place in recent years. This would be a good time, therefore, to take another look at the subject, review the material and circuitry discussed in my previous article, and bring some of it up to date in this month's column.

parameter review

Noise Figure. Noise figures have dropped dramatically over the last few years, largely because of improvements in devices available, but particularly because of the increased availability of inexpensive ($2.50-$15.00) MOS and GaAs FETs. New noise figure measurement equipment that yields extremely accurate numbers has made everyone "more honest" — and while it hasn't yet appeared in many ham shacks, it is readily available at most VHF/UHF conferences.

Modern state-of-the-art preamplifiers can now attain noise figures of 1 dB or better on any Amateur band below 1300 MHz, with gains up to 25 dB, using low-cost (under $15.00) GaAs FETs. In fact, 0.5 dB noise figures are not uncommon when slightly higher priced devices are used and special care is taken to use high "*Q*" components on the input matching networks.

IMD (intermodulation distortion) and blocking. IMD is getting to be a serious VHF/UHF problem as activity, power, and antenna gains are increased. On top of this, it is becoming common practice to place a very-low-noise preamplifier ahead of the receiver to improve sensitivity. Use of GaAs FETs has helped since they usually have high dynamic range and narrower bandwidth than their predecessors, the bipolar transistor, but they frequently have higher gain and hence *increase* the problem! Also, many converters still use mixers with poor dynamic range. To top it off, many VHF/UHF'ers who use converters/transverters have an IF exhibiting poor dynamic range.

How can you cope with these problems? Pay close attention to the gain distribution of the system and keep gain as low as possible ahead of the mixer. Typically speaking, 30 dB of gain ahead of a mixer is usually more than sufficient, even for low-noise and EME. Frequently only 10 to 15 dB of gain is required ahead of a mixer for normal operation such as tropo,* where 2 to 5 dB noise figures are adequate. As a rough rule of thumb, the gain of a preamplifier in a high *dynamic range* receiver should be approximately 6 dB greater than the noise figure of the receiver following it. In a low-noise setup a preamplifier should have about 10 dB more gain than the noise figure of the following receiver. For example, if a converter in a high dynamic range configuration has a 9 dB noise figure, only a 15 dB gain preamplifier is required, but 19 dB would be desired in a *low-noise* system (such as EME). Obviously, if you lower the noise

*Tropospheric communications utilizes weather related changes in the atmosphere as opposed to ion concentrations found in the ionosphere to refract VHF/UHF signals. Using "tropo," reliable communications can be established several hundred miles beyond the horizon. — **Editor.**

figure ahead of the mixer with a moderate noise figure amplifier, less gain is required in the preamplifier. Examples of gain and cascaded noise figure calculations[2] are now found on some computer programs.[3]

Spurious responses. We live in an RF-polluted world. Signals are inundating the entire VHF/UHF spectrum; many of them are not even coming from normal transmitters, but are instead generated by scanners, computers, TV sets, and more recently, CATVI (Community Antenna Television Interference). The days of wide-open (little filtering) front-ends are limited. In order to cope with this situation we must pay more attention to RF filtering, selection of local oscillator frequencies (including the fundamental oscillator when frequency multipliers are used), and use high dynamic range circuitry. When using LO (local oscillator) multipliers for the higher frequency bands, try to use doublers wherever possible. Triplers and quadruplers have all kinds of problems including low output and more spurious products to be filtered. You'll be way ahead in the long run if you don't use them.

IF selection. Let me first reiterate some of the highlights of the earlier material about IF selection.[1] Try to use an IF frequency that is high enough to allow good image rejection but low enough to have good frequency stability. I prefer 28 to 30 MHz and use this range for all my converters through 2304 MHz. For converters from 144 MHz and above, use local oscillators with overtone crystals preferably in the 94 to 116 MHz range. Using a local oscillator with a fundamental frequency of 38.666 or 58 MHz for a 116 MHz local oscillator injection strip for 2 meters (28 MHz IF) is an open invitation for birdies. Furthermore, the lower cost of the crystal is often offset by the cost of the components in the extra multiplier required.

One other recommendation is to *not use even frequencies for the IF*. For example, it is common practice in commercial converters and transverters to use a 404 MHz LO for 432 MHz operation with a 28 MHz IF. This puts the weak signal region (432.0-432.1) between 28 and 28.1 MHz. This is a heavily used frequency range for HF, and IF leakthrough may place some HF signals right on top of a weak signal. If the local oscillator is slightly high in frequency (while still being well within specification), 432.0 MHz signals may be below the tuning range on the IF.

It is reassuring to be able to check frequency calibration accurately with an external frequency marker, but if the marker is a harmonic of a 1, 2, or 4 MHz calibration standard, the image frequency (376.0 MHz in this case), as well as the IF receiver (28 MHz), will pick up the marker. The net result will be a hopeless grouping of signals which must be sorted out before true frequency calibration can be determined. For best results, use an LO that will place the lowest frequency of interest at, for example, 28.1 MHz. (In this case the proper choice of the local oscillator would be 403.9 MHz.) The net result will be a cleaner sounding converter more removed from congestion and only one crystal clear marker to zero beat.

transceiver review

Commercial Amateur transceivers are now available for all VHF/UHF bands up to 1300 MHz. If you have one of these transceivers, there isn't much you can do to the innards without risking possible devaluation if you should ever decide to sell it. The modern rigs are complex and compactly constructed requiring skill, knowledge and complete documentation by anyone attempting to work on them. Real improvements — such as adding a low-noise preamplifier ahead of the receiver — usually have to be made externally.

An external preamplifier, especially a well designed GaAs FET type, will almost always yield a lower noise figure on an existing transceiver. However, most present-day commercial transceivers have low dynamic range "as is" and can generate IMD when any extra gain is placed ahead of them. This is not meant to imply that you won't have the same problem on a homebrew transverter or converter, as discussed earlier. However, in the latter case you will probably be able to lower the gain ahead of the mixer and at least partially compensate for the increase in gain of the extra preamplifier.

Another problem in modern solid-state equipment that has been plaguing Amateurs and commercial users alike is phase noise or noise sidebands present on the local oscillator. This is particularly true on rigs that use synthesized local oscillators. While this problem is not too obvious when listening, when a strong signal appears alongside or sometimes even some distance down the band from the station you are listening to, watch out! Before you tell said station that he has a dirty signal or is hitting it too hard, bypass your preamplifier or turn on the internal attenuator (if you have one) and see whether there is a dramatic drop in QRM or buckshot. Even if you build your own converter or transverter, your IF system can be a limiting factor in dynamic range. You may still have problems similar to the ones mentioned above, but now they may instead appear in the IF circuitry!

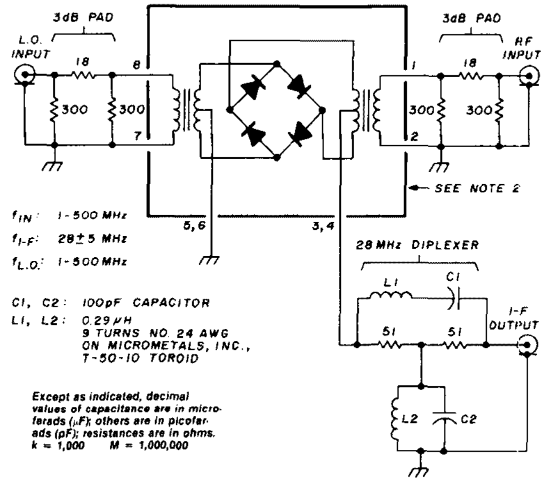
recommended circuits

Mixers. Many of the problems mentioned above can be eliminated or contained by using inexpensive ($10.00 or less) DBMs (double balanced mixers). My low-cost favorite is the Mini-Circuits Labs SRA-1. Their less expensive ($3.95 in quantities of 10 to 49) SBL-1 is also acceptable, but sometimes has a 1 to 3 dB poorer signal handling capability than the SRA-1. Recently I have seen many acceptable DBMs showing up at flea markets for some very attractive prices.

In order to use all of the capabilities of the DBMs, it's important to have each port properly terminated[4] and to provide adequate local oscillator power (5 to 10 milliwatts) at the mixer terminals. I have found that the easiest way to accomplish this is to use 3 dB attenuator pads on the local oscillator and RF ports and a simple diplexer on the IF port. The 3 dB pads will terminate the various undesired frequencies generated internally in the DBM and improve the impedance match to externally connected circuits. The diplexer will filter undesired outputs from the IF while providing a good match to the mixer and postamplifier. A

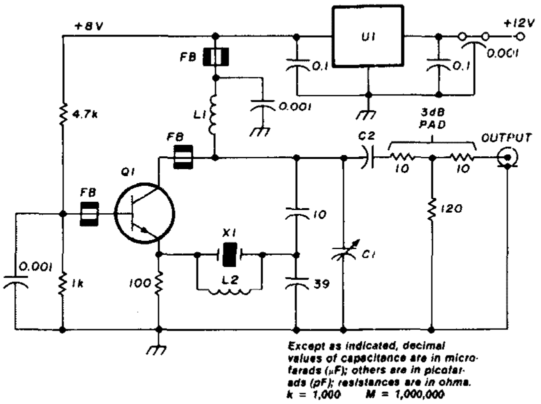
recommended circuit I've been using for over 10 years in shown in **fig. 1**. This circuit required 10 milliwatts of local oscillator power and will easily yield a 9 to 10 dB noise figure as is if followed by a low noise (1 dB) postamplifier. This mixer configuration with a single U310 JFET preamplifier ahead of same is more than adequate for a 2 to 3 dB noise figure high-dynamic-range converter on any Amateur band between 50 and 450 MHz.

Local oscillators. Remember that the local oscillator is also very important. It should be stable, have sufficient output power, and be free of strong (less than 25 dB down) harmonics and any spurious frequenices. My favorite oscillator circuit, an improved version of the one in my 1976 article, is shown in **fig. 2**. Several points still apply. The crystal should be a high-quality 5th or 7th overtone series-resonant type, preferably in an HC-18/U holder. *Capacitors or inductors should never be placed in series with the crystal* because this will dramatically lower the Q of the crystal and hence the stability of the oscillator. If you purchase a high quality, high accuracy crystal, it should be close to the desired frequency. If frequency deviations are needed, compensate for them by offsetting the frequency of the IF. One final point: a common crystal oscillator I sometimes see in use has the crystal connected from base to ground. This is an example of a poorly designed circuit because the crystal is operating in a low Q mode. Hence the oscillator is more likely to be less stable, operate on spurious frequencies, and have high phase noise. If you use such a circuit, try modifying it according to the circuit shown in **fig. 2**.



Notes:
(1) Keep all leads, especially grounds, as short as possible.
(2) Pin connections shown are for Mini-Circuit Labs SRA-1 or SBL-1 DBM. Other manufacturers' DBMs may use a different numbering system, but are usually physically located at the same place on a functional basis.

fig. 1. Recommended circuit for using commercial DBMs.

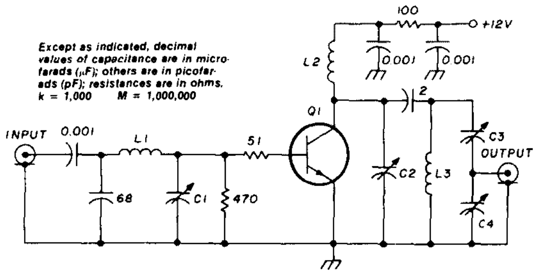


C1	2-20 pF small trimmer capacitor
C2	5 pF capacitor
FB	ferrite bead - Ferroxcube No. 56 590 65/3B or equivalent
L1	8 turns No. 24 AWG enamelled wire close wound on 0.1 inch (2.5 mm) diameter
L2	0.39 μH RF choke or 10 turns No. 28 AWG on T-25-6 toroid core
Q1	2N5179, NE73432 or equivalent
U1	78L08 3-terminal 8-volt regulator
X1	5th or 7th overtone series resonant 90-125 MHz crystal (see text)

fig. 2. Recommended overtone crystal oscillator for 90-125 MHz. Output is approximately 10 milliwatts.

Multipliers. Some typical frequency doublers — improved versions of those illustrated in my earlier article — are shown in **figs. 3** and **4**. High quality tuning capacitors with short leads should be used so that the output power will be high and the spectrum as clean as possible. Properly duplicated, these circuits do not require a spectrum analyzer for alignment. All that is required is to peak all adjustments for maximum output power.

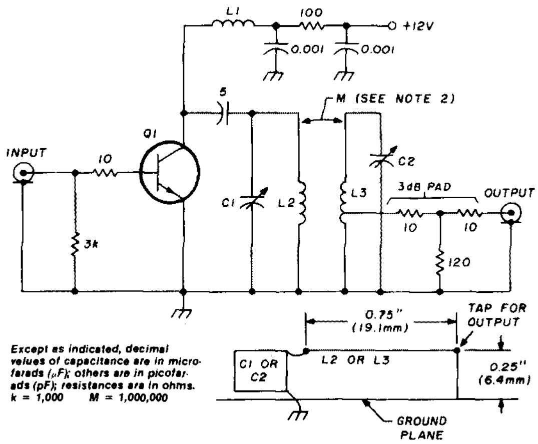
A few extra comments about the multiplier circuits are in order. I prefer to build these circuits with the components soldered or suspended above a piece of double-sided printed circuit board that is attached to the cover of a shielded box such as the Bud CU-124 (or equivalent) with the input/output (for example, BNC) connectors. This technique yields a good ground plane — especially for the bypassing capacitors, tuning capacitors, and inductors — while shielding the circuits from stray pickup. It also allows the unit to be easily connected to



Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

C1,C4	5-50 pF mylar or equivalent trimmer capacitor
C2,C3	2-20 pF mylar or equivalent trimmer capacitor
L1	10 turns No. 24 AWG enameled wire close wound on 0.1 inch (2.5 mm) diameter
L2,L3	4 turns No. 24 AWG wire, turns spaced wire diameter on 0.1 inch (2.5 mm) diameter
Q1	2N5179, NE73432 or equivalent

fig. 3. Frequency doubler: with a 95-110 MHz, 5-10 milliwatt input, a 190-220 MHz, 10-20 milliwatt output is produced.



Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

(1) This circuit is designed without input matching and is to be attached directly to output of doubler in fig. 3 . (See text).
(2) L2 and L3 run parallel and are spaced 0.75 inch (19 mm) apart.

C1,C2	1-10 pF mylar or equivalent trimmer capacitor
L1	10 turns No. 24 AWG enameled wire close wound on 0.1 inch (2.5 mm) diameter
L2	1.0 inch (25 mm) No. 20 AWG spaced 0.25 inch (6.4 mm) above ground (see sketch and Note 2)
L3	same as L2. Tap 0.25 inch (6.4 mm) from ground
Q1	2N5179, NE73432 or equivalent

fig. 4. Frequency doubler: with a 190-220 MHz, 5-10 milliwatt input, a 380-440 MHz, 10-20 milliwatt output is produced.

the mixer with a short coaxial cable and facilitates testing if a spectrum analyzer is available.

The multiplier transistor choice will have a large effect on gain and hence the output power. Modern UHF transistors in TO-92 packages (such as the NEC NE73432 or the Fairchild FMT-1100, if available) are better and will have greater output than the older 2N5179. Also, as noted in **fig. 4**, if you use the second doubler, connect it directly after the circuit shown in **fig. 3**. The output tuning in the first multiplier performs the proper impedance matching.

testing

The easiest test of how your system is performing is to listen on the air, especially during activity nights or during contests. Sensitivity, generally related to noise figure, can be roughly estimated by listening for a distant station. Noise figures can often be tested and optimized at a VHF/UHF conference where noise figure meters are available. *Do not be tempted to retweak the input circuit in your low-noise preamplifier after it has been properly optimized on a good noise figure generator.* Optimum noise figure may frequently yield a lower gain preamplifier, and retweaking input circuits for more output when installed ahead of a converter may seriously degrade the overall noise figure.

You should also listen for unexpected spurious frequencies. Testing is best facilitated if you build your circuits in separate boxes or modules, a technique I have been advocating for many years. This will allow you to have your preamps and/or local oscillator chain easily tested if you have access to a noise figure meter or spectrum analyzer. It will also facilitate a rapid change if a device fails or if you want to substitute a new — hopefully improved — circuit.

final comments

Building your own receiving gear can be quite satisfying and one of the few ways we can get maximum performance with minimum compromise. In the months ahead, more details on high dynamic range and low noise figure preamplifiers will be forthcoming. Hopefully this column will inspire you to get out the soldering iron and put those devices you've been saving to work. In this fast-changing world, the devices we have today may be obsolete tomorrow. . .so let's use and enjoy them today instead of letting them gather dust in a corner of your work bench!

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VHF/UHF coming events
EME perigee weekend: March 17/18.

ham radio

key to 3-element Yagi design

Find driving-point impedance, currents, gain and f/b using pocket calculator

During recent years an interesting series of articles on the Yagi-Uda dipole array antenna were written by the late James Lawson, W2PV.[1] In his first article of the series, he discussed the development of a practical Yagi array and briefly described its electrical parameters. To evaluate the system he employed a computer using a Fortran program which resulted in tabulated values and graphs.

But even after reading Lawson's evaluations and analyses of the Yagi, one intriguing question remained in my mind: could a design or evaluation be accomplished without a computer, using only a handheld programmable calculator?

The answer is *yes*. Such a project can be accomplished using a programmable calculator such as the Hewlett-Packard HP-41C or any non-programmable pocket calculator capable of performing conversions of complex numbers in either rectangular or polar forms. Using any of these calculators, it is possible for the Radio Amateur to determine a Yagi's input impedance, current ratio between dipole elements, free space radiation pattern, forward gain, front-to-back ratio, and the radiation pattern of the total array over smooth earth.

This article addresses the ways in which these parameters can be calculated using a step-by-step procedure. Although a program is provided,* it is recommended that the chapters on alternating currents and vectors in Nelson M. Cooke's *Basic Mathematics for Electronics*[2] (or any similar mathematical text) be reviewed by the reader before beginning the project.

*NOTE: Send SASE to *ham radio*, Greenville, NH 03048.

self-impedance of a dipole

Though several different expressions have been used to characterize a dipole's self-impedance,[3,4] one that is particularly simple to solve on a hand calculator is reproduced below.[5,6]

$$Z_{in} = [122.65 - 204.1\,bl + 110(bl)^2]$$
$$-j\left\{120\left[\log\left(\frac{2l}{a}\right) - 1\right]\cot bl - 162.5 + 140bl - 40(bl)^2\right\} \quad (1)$$

where l = dipole length (normally close to half-wavelength long)
bl = "length" in radians of one leg of the dipole
a = radius of dipole element (same units as "l"

The determination of self-impedance for each Yagi (dipole) element is important because it is part of the total design calculation. Beam elements are assumed to be cylindrical in shape without any taper; elements exhibiting a taper have different current distributions with different input impedances. However, if the taper of the element is gradual, the values given in **table 1** can be used. Thick diameter dipole elements are resonant at lengths shorter than a (physical) half wavelength. The elements of a three-element Yagi are usually standardized in length with the driven element 0.475 wavelength long, the reflector 0.5 wavelength, and the director 0.450 wavelength. The element diameter is based on a size that is structurally sound and large enough to present a low Q — i.e. a slow reactance change with frequency.

The self-impedances and dimensions of a three-element Yagi tuned to 14.15 MHz are listed in **table 1**.

By Walter J. Schulz, Jr., K3OQF, 3617 Nanton Terrace, Philadelphia, Pennsylvania 19154

table 1. Self impedance of dipole elements of different lengths and diameters.

length	radian	1.5 inch OD R	1.5 inch OD X	1.0 inch OD R	1.0 inch OD X
0.500	1.5708	73.0	+j41.0	73.0	+j41.3
0.494	1.5519	70.8	+j28.0	70.8	+j27.0
0.488	1.5331	68.3	+j14.8	68.3	+j12.9
0.484	1.5205	66.6	+j 6.0	66.6	+j 3.5
0.478	1.5017	64.2	−j 7.1	64.2	−j10.5
0.476	1.4954	63.4	−j11.5	63.4	−j15.2
0.475	1.4923	63.0	−j13.6	63.0	−j17.5
0.472	1.4828	61.9	−j20.2	61.9	−j24.5
0.466	1.4640	59.6	−j33.2	59.6	−j28.4
0.462	1.4514	58.1	−j41.9	58.1	−j47.7
0.456	1.4326	56.0	−j54.8	56.0	−j61.6
0.450	1.4137	54.0	−j67.8	54.0	−j75.5
0.444	1.3949	52.0	−j80.7	52.0	−j89.3

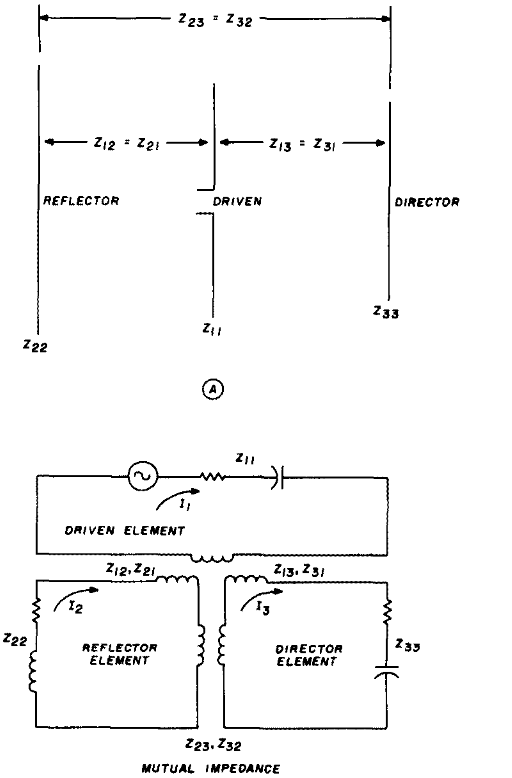

fig. 1A. Three-element Yagi antenna showing self- and mutual-impedance locations. fig. 1B. Network equivalent of three-element Yagi-Uda antenna.

dimension	director	driven	reflector
element diameter	1.5″ (38 mm)	1.5″	1.5″
length (full element)	31.3′ (9.54m)	33′ (10.06m)	34.8′ (10.61m)
length (one leg)	1.41 radians	1.49 radians	1.57 radians

mutual impedances between dipole elements

Mutual impedance is a term that relates current in one element to current in a different element of the same antenna. It's easy to understand if you consider the antenna as a circuit with several meshes or loops. In a circuit, if a voltage is generated in one branch, called a mesh, currents flow in other meshes. The coupling between meshes is through a transfer or mutual impedance common to both.

Many times a T network is used to explain mutual impedance action between two meshes. The T network shunt element is analogous to the mutual impedance displayed in a two-element Yagi antenna. Three-element Yagis require another type of illustration to show circuit relationships.

The circuit equivalent of a Yagi's self- and mutual-impedances are shown in **fig. 1**. Self-impedances are coupled to the other meshes by air core transformers. The transformers have a 1:1 ratio and are assumed lossless. The currents, both in magnitude and phase, can be determined using standard network techniques for each mesh. This is accomplished by writing the simultaneous equations that describe the electrical steady-state condition existing in the network.

$$\begin{array}{ll} \text{driven element} & I = I_1Z_{11} + I_2Z_{12} + I_3Z_{13} \\ \text{reflector element} & 0 = I_1Z_{21} + I_2Z_{22} + I_3Z_{23} \\ \text{director element} & 0 = I_1Z_{31} + I_2Z_{32} + I_3Z_{33} \end{array} \quad \textbf{(2)}$$

where I_1, I_2, and I_3 are the currents that flow in the driven, reflector, and director elements, respectively, and

Z_{11}, Z_{22}, Z_{33} are the same elements' self-impedances;

Z_{12}, Z_{13}, Z_{21}, Z_{23}, Z_{31}, and Z_{32} are mutual impedances between subscripted elements.

For example, Z_{12} = the mutual impedance between elements 1 and 2.

A list of mutual impedances[1] for different element spacing is given in table 2.

Notice from eq. 2 how current magnitude and phase in each element is controlled by its self- and mutual-impedances. Once the element length and diameter is chosen, self-impedance becomes a fixed value. Therefore, the only means of controlling the current magnitude and phase in each element is by the mutual impedance values. Mutual impedance values change when different physical spacings between elements are used. The greater the element spacing, the less effect the mutual impedance has on the driving-point (input) impedance of the antenna. Consequently, changing the spacing changes the mutual impedances, which change the current ratio between elements. These currents in turn determine the radiation pattern of the antenna (gain and f/b). In a *driven* vertical array the current phase of each element is controlled by a phase delay line or network while the *parasitic* Yagi antenna relies on element spacing and length to diameter ratio to control current phase through self- and mutual impedances.

The solution to the unknown currents flowing in each of the three meshes is found by using determinants.

driven element
$$I_1 = \frac{(Z_{22}Z_{33}) - (Z_{32})^2}{\Delta}$$

reflector element
$$I_2 = \frac{(Z_{23}Z_{31}) - (Z_{33}Z_{21})}{\Delta} \qquad \textbf{(3A)}$$

director element
$$I_3 = \frac{(Z_{21}Z_{32}) - (Z_{31}Z_{22})}{\Delta}$$

$$\Delta = (Z_{11}Z_{22}Z_{33} + (Z_{12}Z_{23}Z_{31}) + (Z_{13}Z_{21}Z_{32}) - (Z_{31}Z_{22}Z_{13}) - (Z_{32}Z_{23}Z_{11}) - (Z_{33}Z_{21}Z_{12}) \qquad \textbf{(3B)}$$

The three currents may be given in rectangular form but it is more helpful to express them in polar form, because the latter shows whether the current phase is leading or lagging. Note this is very helpful to check to see if the solutions are correct. The reflector current phase should be positive (leading) while the director current phase should be negative (lagging). Each of the current characteristics denotes that the parasitic elements are either inductive or capacitive reactive.

calculating the driving-point impedance

One of the reasons for calculating currents in each element is to determine the driving-point impedance at the driven element of the Yagi antenna. Knowing this impedance, one can now match it to the transmission line. Wide spacing between elements usually produces higher driving-point resistance (first term of the complex impedance). This results in lower Q and wider bandwidths. The driving-point impedance equals:

$$Z_{in} = Z_{11} + (I_2/I_1)Z_{12} + (I_3/I_1)Z_{13} \qquad \textbf{(4)}$$

table 2. Mutual impedances between two elements.

spacing (in wavelengths)	R	X*
0.00	73	42
0.05	72	24
0.10	67	7
0.15	60	− 7
0.20	51	−19
0.25	41	−28
0.30	29	−34
0.35	17	−37
0.40	6	−37
0.45	− 4	−35
0.50	−12	−30
0.55	−19	−23
0.60	−23	−16
0.65	−25	− 8
0.70	−25	−0.25
0.75	−22	6
0.80	−18	12
0.85	−13	16
0.90	− 7	18
0.95	− 1	19
1.00	4.0	18
1.05	9	15
1.10	12	11
1.15	14	7
1.20	15	2
1.25	14	− 3
1.30	12	− 7
1.35	10	−10

A three-element array is to be constructed with a reflector to driven element spacing of 0.1 wavelength and a director to driven element spacing of 0.15 wavelength. The element diameters are 1.5 inch and their self- and mutual-impedance values are taken from tables 1 and 2, respectively:

self impedances	mutual impedances
$Z_{11} = 63 - j15$	$Z_{12} = Z_{21} = 67 + j7$
$Z_{22} = 73 + j41$	$Z_{31} = Z_{13} = 60 - j7$
$Z_{33} = 54 - j70$	$Z_{23} = Z_{32} = 41 - j28$

Insert these values into eqs. 3A and 3B to solve for the currents and then use eq. 4 to solve for Z_{in}. This can be done manually or by use of a calculator. I used an HP-41C and a quad memory module. If you have this calculator, clear all registers, key in size 100, then key self- and mutual-impedance values into the proper memory storage registers. To obtain current magnitude and phase for each element current ratio and driving-point impedance, execute program "ZZ".

finding the determinant

Using this procedure one obtains the value for Δ and

*Though other sets of values exist, this procedure is the key feature of this article - Ed

the values for driven, reflector, and director element currents I_1, I_2, and I_3 respectively.

$$\Delta = 95476 - j182530$$

driven element

$$I_1 = \frac{(Z_{22}Z_{33}) - (Z_{32})^2}{\Delta}$$

$$= \frac{(73+j41)(54-j70)-(41-j28)^2}{\Delta}$$

$$= 0.0159 + j0.0241$$

reflector element

$$I_2 = \frac{(Z_{23}Z_{31}) - (Z_{33}Z_{21})}{\Delta}$$

$$= \frac{41-j28)(60-j7)-(54-j70)(67+j7)}{\Delta}$$

$$= -0.0142 - j0.0027$$

director element

$$I_3 = \frac{(Z_{21}Z_{32}) - Z_{31}Z_{22})}{\Delta}$$

$$= \frac{(67+j)(41-j28)-(60-j7)(73+j41)}{\Delta}$$

$$= 0.0113 - j0.0154$$

current ratios between elements

$$I_2/I_1 = -0.3484 + j0.3611$$

$$I_3/I_1 = 0.2284 - j0.6213$$

driving-point impedance

These numbers when substituted into **eq. 4** give us the value of the input or driving-point impedance:

$$Z_{in} = Z_{11} + (I_2/I_1)Z_{12} + (I_3/I_1)Z_{13}$$

$$= (63-j15)+(0.3484+j0.3611)(67+j7)$$

$$+ (-0.2284-j0.6213)(60-j7)$$

$$= 19.0753 - j28.9239$$

The method outlined above is the simplest procedure that can be used to find the driving-point impedance of a monoband Yagi-Uda dipole array. It is hoped that this information will prove helpful to those Radio Amateurs considering designing and building their own three-element Yagi-Uda beam antennas on the high frequency Radio Amateur bands.

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Bill Orr W6SAI

broadband 80/160-meter antenna

One of the very interesting advantages of writing this column is the feedback I get from readers. A case in point: in my October, 1983, column I discussed the problem of building a simple broadband antenna that would cover the whole of either the 80- or 160-meter bands with a reasonably low value of SWR on the feedline. Some of the newer solid-state transmitters are quite sensitive to an SWR other than 1.0:1, and they react by reducing the output power of the final amplifier stages at high values of SWR.

One of the antennas I discussed was the crossed-dipole array described by Mason Logan, K4MT, in the May, 1983, issue of this magazine. His basic antenna design is shown in **fig. 1**. The measured SWR curve of this antenna is shown in **fig. 2**. I suggested in my October column that a matching coil might be required at the antenna feedpoint to bring the impedance closer to 50 ohms.

Shortly after publication, I received a note from K4MT stating, in part, "Your statement that the antenna impedance is quite low and that a matching coil across the feedpoint is needed is not correct. With the stagger-tuned dipoles, each dipole acts as the network for the other. . .Nothing more is needed!"

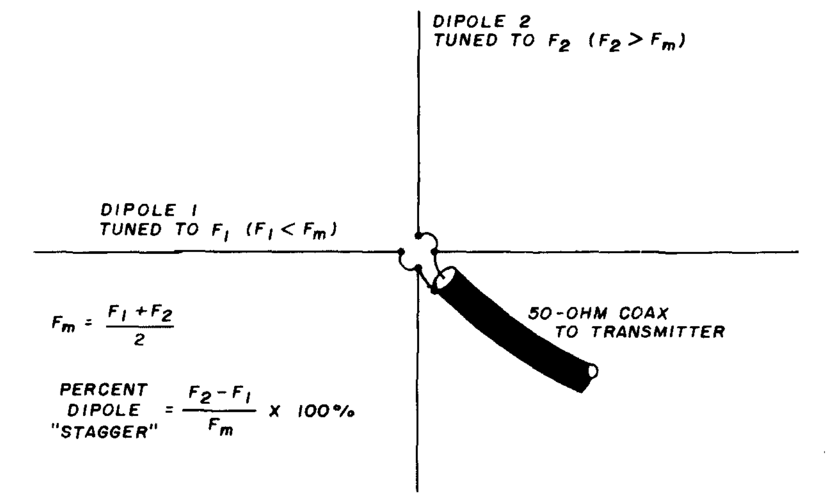


fig. 1. Top view of broadband stagger-tuned, crossed-dipole antenna.

Logan is correct and I am wrong, as his letter proves. He goes on to say that for the stagger-tuned dipoles, between the two chosen resonant frequencies, the reactances of the dipoles have opposite signs, forming a lossy, antiresonant circuit which can have an impedance maximum near the center frequency where the reactances are equal in magnitude. Near the band edges, at the resonant frequencies of the dipoles, the impedance is somewhat less than that of each dipole alone. Hence the W-shaped curve for the impedance as well as the SWR.

Mason goes on to say that height of the antenna above ground has a significant impedance effect and that when the resonant points are properly chosen, a satisfactory SWR curve can be achieved for heights of one-quarter wave or less. Great news for the "top-band" operator!

Paul Scholz, W6PYK, has worked with Mason to develop a computer program that determines the best design frequencies for the crossed-dipole antenna and provides im-

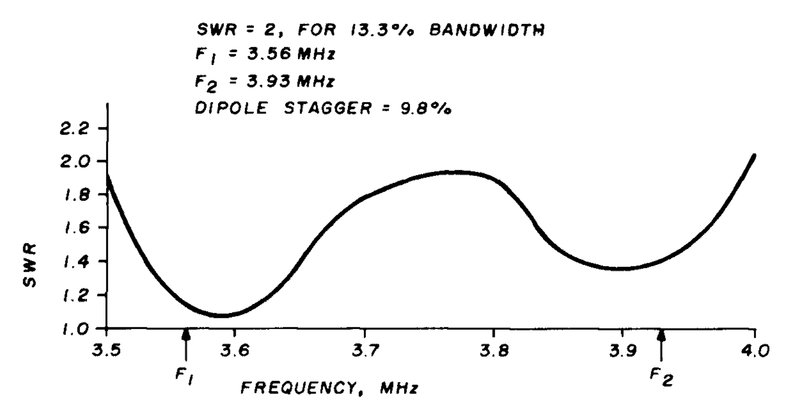


fig. 2. Measured SWR of broadband 80-meter stagger-tuned antenna using inverted Vs.

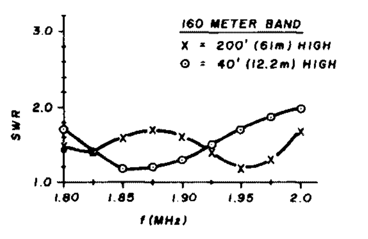


fig. 3. Computer-derived SWR curves for K4MT crossed-dipole antenna for 160-meter band. (Computer program by Paul Scholz, W6PYK.)

pedance and SWR readout. Two examples, given in **fig. 3** and **4**, show that even at low height, both the 80 and 160-meter designs exhibit a good match to a 50-ohm line; a better match, in fact, than if the antenna were suspended higher in the air.

The 160-meter design is summarized in **fig. 3**. The dipoles were cut by formula to 1.75 MHz and 2.1 MHz (outside both ends of the 160-meter band) in the case of the 40-foot high antenna, and to 1.8 MHz and 1.975 MHz in the case of the 200-foot high antenna.

In each case, the resonant frequencies were chosen to provide a satisfactory value of SWR across the band (less than 2 to 1). The 40-foot high configuration is of most interest because it is a practical situation that can be duplicated by the average Amateur.

Only a portion of the "W" shape shows in the curve, as the higher design point was chosen outside the high frequency end of the band. Compare this curve with your ordinary 160-meter dipole located at a 40-foot elevation!

The "W" shape shows up in the 200-foot high antenna as the design points are closer together. But who can place an antenna at the 200-foot level? Not me.

Fig. 4 shows two crossed-dipole SWR curves for the 80-meter band. One antenna is 100 feet high and the design points are 3.55 MHz and 3.9 MHz. Note that the minimum SWR points do not correspond exactly to the design frequencies. The design points of the 40-foot high dipoles are 3.525 MHz and 3.975 MHz. Both of these antennas provide good SWR curves, with the lower antenna especially attractive for everyday operation across the band.

In summary, the K4MT crossed-dipole, broadband antennas do not exhibit critical design requirements and should be trimmed at the specific location for best match.

For those who want to write their own computer program for this antenna, the necessary information is given in **fig. 4** of K4MT's orignial article. (Thanks to K4MT and W6PYK for forwarding the computer data and additional design information to me.)

the K2GNC Y-doublet for 80 meters

Other hams have been experimenting with broadband antennas for 80 and 160 meters: Bill Pfaff, K2GNC, has come up with the interesting concept shown in **fig. 5**. He's had his Y-doublet up for over three years and it's worked quite well. The antenna is supported by a pole in the center, similar to that of an inverted-V. When properly constructed, it covers the entire 80-meter band (3.5 MHz to 4.0 MHz) with a SWR less than 1.5 to 1. This requires proper length and orientation of the three unequal-length elements.

The drawing shows connection of the three antenna wires to the feedline. Two wires (radials?) are attached to the shield of the line and a third antenna wire is attached to the center con-

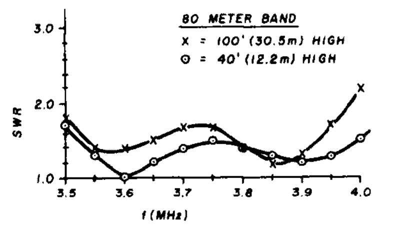


fig. 4. Computer-derived SWR curves for K4MT crossed-dipole antenna for 80-meter band. (Computer program by Paul Scholz, W6PYK.)

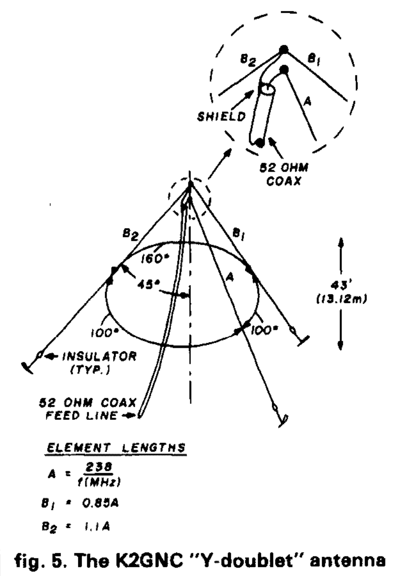


fig. 5. The K2GNC "Y-doublet" antenna for 80 meters. Antenna covers complete band with low SWR. It is supported from a single pole.

ductor. The wires form an angle of 45 degrees with respect to the 45-foot wooden support. The antenna wires also help guy the pole.

The two "radials" (marked B1 and B2 in the drawing) are located 100 degrees away from the radiator (marked A), as viewed in the horizontal plane. Minimum SWR frequencies are controlled by the B1 and B2 wire lengths. K2GNC has adjusted his wires so that the SWR of the antenna is below 1.5 to 1 from 3530 kHz and 3980 kHz. Outside these limits, the SWR rises sharply. SWR at the band edges can be reduced, but at the expense of high mid-band SWR.

design example

Choosing a design frequency of 3800 kHz generates the following element lengths:

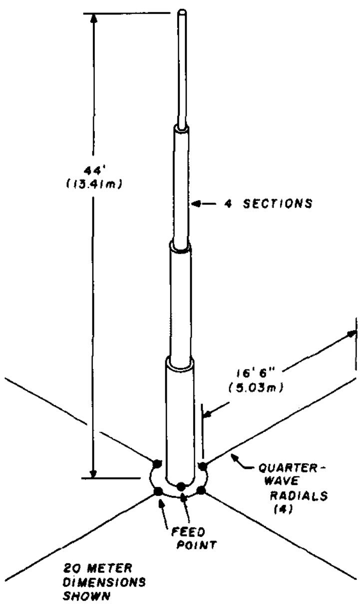
A: 62.63 feet or 62 feet 7 inches (=238/3.8)
B: 53.24 feet or 52 feet 3 inches (=0.85 × 62.63)
C: 68.89 feet or 68 feet 10 inches (=1.1 × 62.63)

The length and height of B1 and B2 can be varied.

A 10-meter model was assembled and placed on a rotator. Field strength measurements revealed a nearly-omnidirectional pattern, with narrow, deep nulls on each side of wires B1 and B2. K2GNC suggests these nulls may be due to the presence of nearby objects.

The feedline should come straight down to the ground underneath the antenna. The use of a balun did not affect measured SWR, nor antenna operation.

I spoke to Bill Pfaff on the phone about this interesting antenna and mentioned the K4MT crossed-dipoles. I asked him if he thought his antenna was a relative of K4MT's, and what would happen if he added a second wire to the A wire, running away from it, making it a four-wire configuration. Bill said he'd tried this idea and found that it didn't work as well as the present designs. So perhaps the K4MT and K2GNC antennas don't have that much in common, after all.



FREQUENCY (MHz)	WHIP	RADIAL
7.0	89' 0" (27.13m)	33' 5" (10.19m)
10.1	61' 9" (18.82m)	23' 2" (7.06m)
14.17	44' 0" (13.41m)	16' 6" (5.03m)
18.11	34' 5" (10.49m)	12' 10" (3.91m)
21.22	29' 4" (8.94m)	11' 0" (3.33m)
24.94	25' 0" (7.62m)	9' 4" (2.84m)
28.6	21' 9" (6.63m)	8' 2" (2.49m)

fig. 6. The 5/8-wave HF vertical antenna. "Whip" is made of four 12-foot telescoping sections of aluminum tubing. Taper ratio from bottom to top is 2:1.

the extended ground plane for HF operation

One of the long-standing jokes about the ground plane antenna is that because it's omnidirectional, it's equally poor in all directions! Maybe so, but an examination of DX QSLs reveals that a large percentage of overseas stations uses ground plane antennas, and some of these signals are quite powerful.

VHF operators have popularized a 5/8-wavelength vertical whip antenna which provides 3 dB gain over a simple 1/4-wave ground plane. It is possible to adapt such an antenna to a lower frequency just as it is done at VHF.

Originally, the 5/8-wave vertical antenna was designed some decades ago for use as an "anti-fade" antenna for the broadcast band.

A representative HF design is shown in **fig. 6**. Dimensions are shown for a center frequency of 14.17 MHz, using tubing for the element. Because the tubing is telescoping, there's a slight taper effect which must be taken into account. The final installation uses a 44-foot vertical section, and many quarter-wave radial wires beneath it. The general formula is:

length (feet) = 623.5/f (MHz)

The antenna must be tuned to resonance, and the easiest way to do this is to add enough inductance at the base to make the overall system resonant at an odd quarter-wavelength mode (three-quarter waves).

Three-quarter wave resonance is determined by adjusting the number of turns in the base coil until a dip meter coupled to the coil-antenna system indicates 14.17 MHz. The bottom end of the coil is attached to the radial wires, which fan out in a horizontal plane.

Once the antenna is resonant, the transmission line is tapped on a few turns above the bottom end of the coil and the tap varied until lowest SWR is achieved (**fig. 7**). It may be necessary to adjust the coil a fraction of a turn to drop the SWR to its lowest possible value.

Amateurs accustomed to the performance of a simple ground plane antenna will find this extended version to be a vastly improved design for both receivng and transmitting.

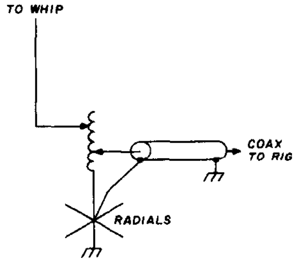


fig. 7. Single coil resonates whip to 3/4-wave mode and also permits match to coaxial line. For 14 MHz, coil is 12 turns of No. 12 wire, spaced twice wire diameter, wound on a ceramic form 2 inches in diameter. Adjust taps for best match as described in text.

table 1. Comparison of Belden 9913 cable with RG-8/U, RG-8A/U and RG-213/U.

f(MHz)	nominal attenuation RG-8/U, RG-8A/U, RG-213/U	(dB/100 feet) 9913
50	1.6	0.9
100	2.2	1.4
200	3.2	1.8
400	4.7	2.6
700	6.9	3.6
900	8.0	4.2
1000	8.9	4.5
4000	21.5	11.0

a coax "sleeper"

Reid Brandon, W6MTF, has pointed out an interesting coaxial cable listed in the new Belden wire and cable catalog. It may be 1984's replacement for the old RG-8/U, RG-8A/U, and RG-213/U. The cable is RG-8/U size, so the fittings for the old cable will work with the new one. It has a solid (not braided) center conductor and 61 percent coverage of the outer braid, plus a conductive, 100 percent coverage solid flexible metallic sheath beneath the braid. The dielectric is called "semi-solid polyethylene," which is not to be confused with foam dielectric. Instead, the new dielectric looks as if it is shaped to provide small air spaces (instead of foam bubbles) along the line. The velocity of propagation is 84 percent as opposed to 66 percent for the RG-8/U type line.

The attenuation of the Belden 9913 cable compared to some of the older varieties is listed in **table 1**. While the virtues of the cable are not apparent below 50 MHz, there's a big payoff in the VHF/UHF region. (Just check the 400 MHz figures!) And the newer cable is a lot less expensive than a power amplifier. For those interested in even less loss in the VHF/UHF region, Belden *9913* is suggested over the readily available forms of RG-8.

ham radio

the HP-IB greatly simplified

Test instruments communicate accurately and rapidly

The term **GPIB** (General Purpose Interface Bus), HP-IB (Hewlett-Packard Interface Bus), and IEEE-488 standard all mean the same thing: they describe a bus-oriented interface for instruments. For the sake of simplicity in this discussion, let's use the term HP-IB since Hewlett-Packard was instrumental in its adoption as an interface standard between lab instruments and calculators or minicomputers.

This article explains the bus concept, defines its signal groupings and individual signals, and provides an historical overview of how specific requirements prompted the development of this concept.

need for standardization

In 1972 Hewlett Packard, along with the United States Digital Instrumentation Committee for Standardization of Interfaces, developed the document that was adopted as a standard two years later by the IEC (International Electrotechnical Commission) by voting on a format ballot.

The decision to use automated, rather than manually controlled systems was based on four main advantages:

- Elimination of operator fatigue, yielding absolutely consistent results on repeated measurements.
- Greater throughput or faster processing speeds.
- More thorough testing resulting from this enhanced speed.
- Results expressed in scientific or engineering notation (i.e., powers of three such as *milli, micro, kilo,* and *mega*).

Table 1 lists key functional needs which are met by the system user who applies the HP-IB to his instrumentation.

how the HP-IB is used

The HP-IB provides a functional and electrical interface for up to sixteen laboratory instruments (DVMs, signal generators, and frequency counters, for example) daisy-chained or linked together in parallel with a controller such as a minicomputer or calculator (**fig. 1**). A controller's function is to designate which devices on the bus will be "listeners" (receivers) and which ones will be "talkers" (senders). The simplest possible system, then, would consist of two devices — a "talker" and a "listener" — without any need for a controller. One device could only "talk" while the other "listened". No interaction, however, would be possible. The maximum allowable distance between instruments is approximately 60 feet for serial digital data transmissions in the 1,000,000 bytes per second range.

Sixteen signal lines are required for the HP-IB. These are broken up into three distinct functional groups: (1) eight data lines; (2) five control lines; and (3) three handshaking lines. These terms will be explained shortly.

These three distinct functional groups are further divided into three component buses. A bus can be best thought of as a conduit through which only one type of signal flows. As an example, a control signal would never flow on a data bus and vice-versa (see **fig. 2**).

The eight bidirectional data bus lines are DI0 through DI7. These carry seven bits of coded interface messages in ASCII format and device-dependent

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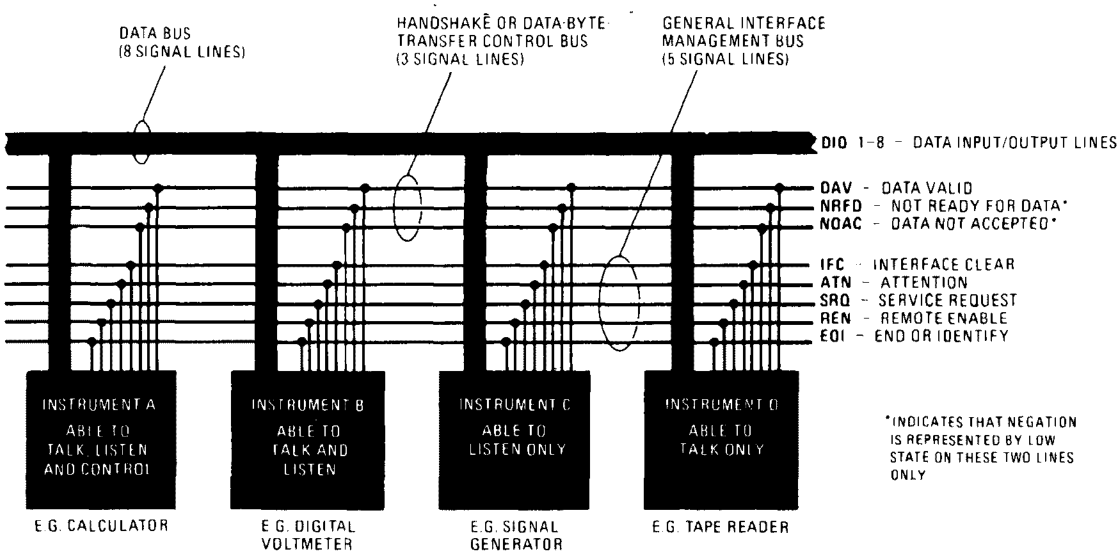


fig. 1. The HP-IB bus uses a 16-line cable to quickly link any instruments equipped with appropriate interface circuitry into a system. Data transfer is byte serial, bit-parallel at rates as high as 1 megabyte per second.

table 1. Key functional needs accommodated by HP-IB (IEEE 488).

basic need	concept	capability provided by
Unambiguous definition	Logical definition independent of implementation scheme	State diagram description of interface functions
Direct access to multiple asynchronous messages	Dedicated signal lines	IFC, ATN, SRQ, REN, EOI signal lines
Cost/performance flexibility	Optionality	Ten interface functions with allowable subsets
Multiple listeners independent of position or response rate	Three-wire handshake	DAV, NRFD, NDAC signal lines
Minimal hardware cost	Bus structure with minimal signal line count	Bi-directional bus for address, command, data, status messages
Standard method for accessing devices	Common code, easily generated and used	Address and universal command structure based on ASCII code
Slow speed status reporting	Device initiated service request	Common SRQ signal line, Serial Poll Mode with status byte reporting
High speed status reporting	Controller initiated status request	Parallel Poll Mode, one status bit for each of eight devices
Accommodation of other interface techniques	Hierarchical partitions	Terminal unit dedicated to interface conversion for cluster of local devices

messages from one to eight bits. But first, let's look at ASCII. **Table 2** shows what ASCII (American Standard Code for Information Interchange) codes are. Note that not all are printable or displayable; some are control characters — that is, carriage return, paper advance, backspace, etc.

The three handshaking signals are really signals that ask a question such as, "Is the serial transmission complete?" The queried device then answers with "yes" or "no." They are the status signals that give the box what appears to be an ability to think; therefore, many of these boxes are called "smart" boxes.

These three handshaking or interface lines (DAV, NRFD, and NDAC) effect the transfer of each byte of data on the data bus from an addressed talker (sender) to all addressed listeners (receivers). The source device,

table 2. ASCII codes and characters.

ASCII & IEEE 488 (GPIB) CODE CHART

B4	B3	B2	B1	Ø Ø Ø	Ø Ø 1	Ø 1 Ø	Ø 1 1	1 Ø Ø	1 Ø 1	1 1 Ø	1 1 1
BITS			B7 B6 B5	CONTROL		NUMBERS SYMBOLS		UPPER CASE		LOWER	
Ø	Ø	Ø	Ø	0 NUL 0 (0)	20 DLE 10 (16)	40 SP 20 (32)	60 0 30 (48)	100 @ 40 (64)	120 P 50 (80)	140 ` 60 (96)	160 p 70 (112)
Ø	Ø	Ø	1	1 GTL SOH 1 (1)	21 LLO DC1 11 (17)	41 ! 21 (33)	61 1 31 (49)	101 A 41 (65)	121 Q 51 (81)	141 a 61 (97)	161 q 71 (113)
Ø	Ø	1	Ø	2 STX 2 (2)	22 DC2 12 (18)	42 " 22 (34)	62 2 32 (50)	102 B 42 (66)	122 R 52 (82)	142 b 62 (98)	162 r 72 (114)
Ø	Ø	1	1	3 ETX 3 (3)	23 DC3 13 (19)	43 # 23 (35)	63 3 33 (51)	103 C 43 (67)	123 S 53 (83)	143 c 63 (99)	163 s 73 (115)
Ø	1	Ø	Ø	4 SDC EOT 4 (4)	24 DCL DC4 14 (20)	44 $ 24 (36)	64 4 34 (52)	104 D 44 (68)	124 T 54 (84)	144 d 64 (100)	164 t 74 (116)
Ø	1	Ø	1	5 PPC ENQ 5 (5)	25 PPU NAK 15 (21)	45 % 25 (37)	65 5 35 (53)	105 E 45 (69)	125 U 55 (85)	145 e 65 (101)	165 u 75 (117)
Ø	1	1	Ø	6 ACK 6 (6)	26 SYN 16 (22)	46 & 26 (38)	66 6 36 (54)	106 F 46 (70)	126 V 56 (86)	146 f 66 (102)	166 v 76 (118)
Ø	1	1	1	7 BEL 7 (7)	27 ETB 17 (23)	47 ' 27 (39)	67 7 37 (55)	107 G 47 (71)	127 W 57 (87)	147 g 67 (103)	167 w 77 (119)
1	Ø	Ø	Ø	10 GET BS 8 (8)	30 SPE CAN 18 (24)	50 (28 (40)	70 8 38 (56)	110 H 48 (72)	130 X 58 (88)	150 h 68 (104)	170 x 78 (120)
1	Ø	Ø	1	11 TCT HT 9 (9)	31 SPD EM 19 (25)	51) 29 (41)	71 9 39 (57)	111 I 49 (73)	131 Y 59 (89)	151 i 69 (105)	171 y 79 (121)
1	Ø	1	Ø	12 LF A (10)	32 SUB 1A (26)	52 * 2A (42)	72 : 3A (58)	112 J 4A (74)	132 Z 5A (90)	152 j 6A (106)	172 z 7A (122)
1	Ø	1	1	13 VT B (11)	33 ESC 1B (27)	53 + 2B (43)	73 ; 3B (59)	113 K 4B (75)	133 [5B (91)	153 k 6B (107)	173 { 7B (123)
1	1	Ø	Ø	14 FF C (12)	34 FS 1C (28)	54 , 2C (44)	74 < 3C (60)	114 L 4C (76)	134 \ 5C (92)	154 l 6C (108)	174 | 7C (124)
1	1	Ø	1	15 CR D (13)	35 GS 1D (29)	55 – 2D (45)	75 = 3D (61)	115 M 4D (77)	135] 5D (93)	155 m 6D (109)	175 } 7D (125)
1	1	1	Ø	16 SO E (14)	36 RS 1E (30)	56 . 2E (46)	76 > 3E (62)	116 N 4E (78)	136 ^ 5E (94)	156 n 6E (110)	176 ~ 7E (126)
1	1	1	1	17 SI F (15)	37 US 1F (31)	57 / 2F (47)	77 UNL ? 3F (63)	117 O 4F (79)	137 UNT _ 5F (95)	157 o 6F (111)	177 RUBOUT (DEL) 7F (127)

ADDRESSED COMMANDS (columns Ø Ø Ø) · UNIVERSAL COMMANDS (columns Ø Ø 1) · LISTEN ADDRESSES (columns Ø 1 Ø, Ø 1 1) · TALK ADDRESSES (columns 1 Ø Ø, 1 Ø 1) · SECONDARY ADDRESSES OR COMMANDS (columns 1 1 Ø, 1 1 1)

KEY TO CHART

octal — 25	PPU — GPIB code	
	NAK — ASCII character	
hex — 15	(21) — decimal	

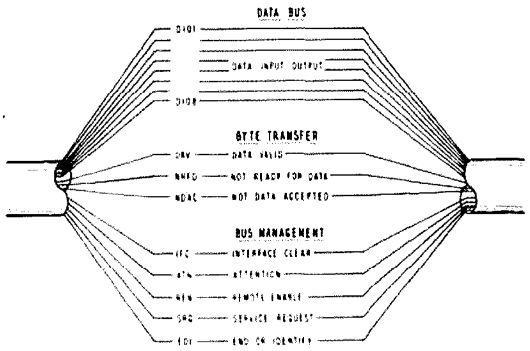


fig. 2. Party-line bus structure services 15 devices.

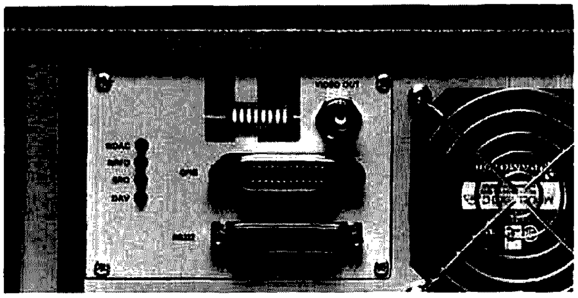


fig. 3. Rear panel switches allow HP-IB to be addressed by controller or operate in "talk only" mode.

whether controller or talker, drives the DAV (Data Valid) line which indicates the availability and validity of information on the data bus, while the acceptors (listeners) drive the NRFD (Not Ready For Data) and the NDAC (Not Data Accepted) lines.

Instruments often possess rear panel switches (**fig. 3**) that allow particular instruments to be addressable by a controller or operate simply in a "talk only" mode to another device such as a printer. If rear panel switches are not present, this switching function is probably accomplished by PC board jumper wires.

Five control or general management signal lines control the orderly flow of information across the interface as follows:

- The ATN (attention) line differentiates between data and control messages on the DIO lines.
- The EOI (End Or Identify) code indicates completion of a multiple byte transfer sequence and can also, together with ATN, activate a Parallel Poll.
- The IFC (Interface Clear) enables the interface system, placing all devices in a known quiet state prior to executing a bus transaction.
- The REN (Remote Enable) line selects between remote or local sources of device programming data.
- The SRQ (Service Request) is a call for service from one of the devices on the bus to the controller.

The three-wire (interlocked) handshake transfer scheme (DAV, NRFD, and NDAC) described above utilizes byte-serial, bit-parallel data travel on the eight DIO lines at typical speeds of 200 to 250 kilobytes per second (**fig. 4**). The maximum data transmission rate is 1 megabyte over extremely limited distances. The faster the data transmission, the shorter the allowable interconnecting cables can be; however, there is one way to overcome this limitation; see **fig. 5**. The HP37201A HP-IB Extender solves this problem by converting parallel data from the interface bus into a serial bit stream. This bit stream is suitable for transmission to a remote site

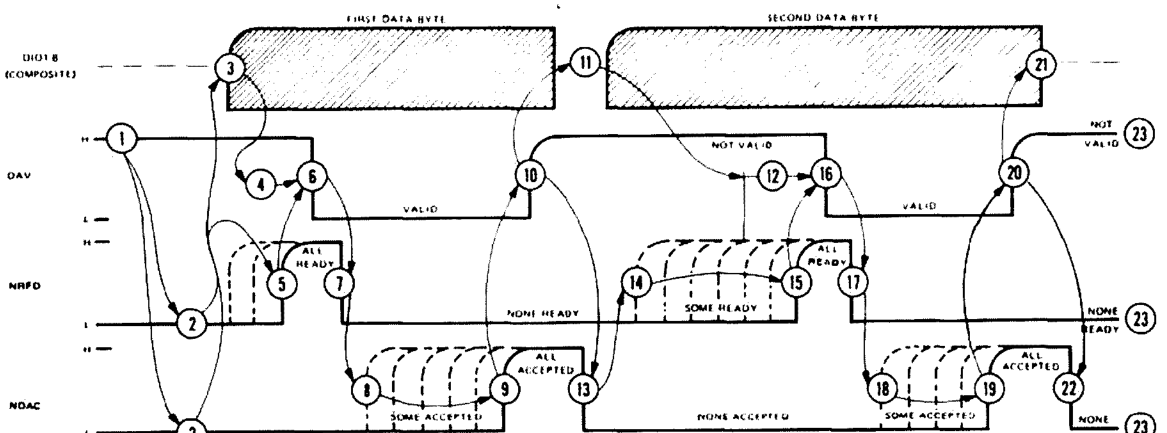


fig. 4. The interlocked handshake transfers data bytes asynchronously and unambiguously to one or more listeners.

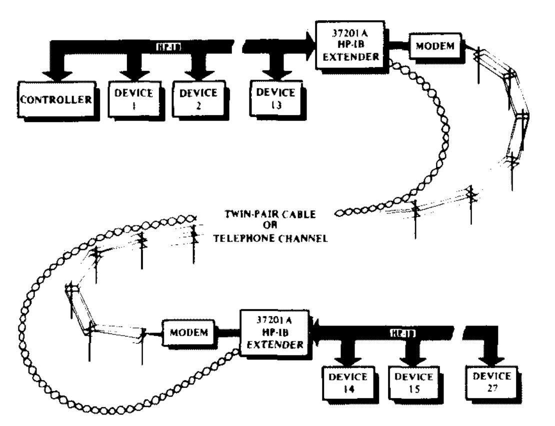


fig. 5. Point-to-point connection using twin twisted pair cable or full duplex modem link.

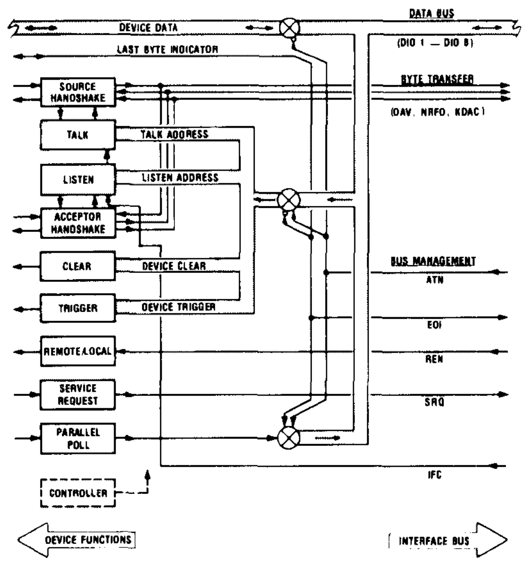


fig. 6. Ten interface functions serve to link the bus structure to the device functions.

and reconverts incoming serial data into the bit-parallel HP-IB format. This allows an HP-IB system to be split into two or more discrete parts separated by HP-IB extenders and a serial data link. A range of 0.6 mile (1 km) is possible if twin-pair cables are used for the transmission path, and virtually unlimited range is possible with a telephone modem hookup. Full duplex operation is also possible and allows for data to flow in both directions at the same time. Many HP-IB problems stem from glitches in this handshaking process. In part, this crisis results because the data transfer is asynchronous. A number of bus analyzers offer some means of detecting protocol violations in handshake activity.

The Hewlett-Packard Interface Bus, HP-IB INTERFACE functions (**fig. 6**) include:

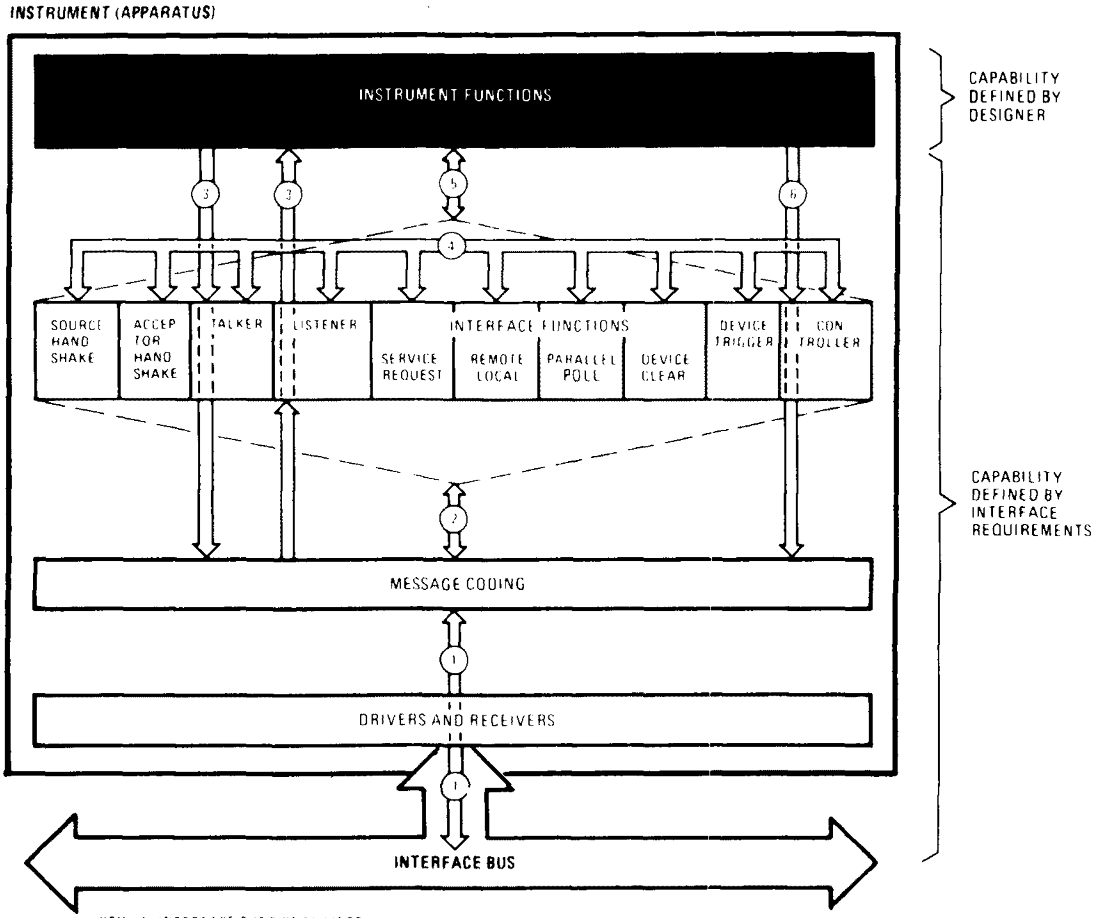
- Talker
- Listener
- Controller
- Source Handshake
- Acceptor Handshake
- SR or Service Report
- RL or Remote Local
- PP or Parallel Poll
- DC or Device Clear
- DT or Device Trigger

The first five functions exist in virtually every HP-IB System as primary functions. The last five, less frequently encountered, are found on the Hewlett-Packard Interface Bus. The detailed functions of these last five are defined as follows:

- **SR (SERVICE REPORT)** allows the device to ask the controller for attention asynchronously in relation to other messages.
- **RL (REMOTE LOCAL)** determines whether messages from the data bus (remote program) or from some local program source will receive the program data from a device in the system.
- **PP (PARALLEL POLL)** is a rapid simultaneous status report to the controller from several devices.
- **DC (DEVICE CLEAR)** enables an instrument (IFC enables the interface itself).
- **DT (DEVICE TRIGGER)** initiates a user-specified transaction within device functions and will synchronize similar actions for multiple measurements among a number of devices after programming.

For designers tackling the problem, the first design specification written is in terms of interface functions (not instrument functions), messages to and from the interface, and the behavior of each of these functions. This last item is traditionally accomplished with state diagrams in which signal flows and events happening are "tied together". **Fig. 7**, which probably best illustrates this process, will not only give you a feel for what goes into the process, but also a fuller appreciation of this clever interfacing scheme. Lastly, this is admittedly a complicated subject and if all of it is not totally clear, don't become frustrated. Once you begin working with them, the concepts will become second nature to you.

This has been an overview or summary of the HP-IB which you are certain to see much more of in the immediate future if you work with modern test equipment at all.



KEY
1 INTERFACE BUS SIGNAL LINES
2 REMOTE INTERFACE MESSAGES TO AND FROM INTERFACE FUNCTIONS
3 INSTRUMENT DEPENDENT MESSAGES TO AND FROM INSTRUMENT FUNCTIONS
4 STATE LINKAGES BETWEEN INTERFACE FUNCTIONS
5 LOCAL MESSAGES BETWEEN INSTRUMENT FUNCTIONS AND INTERFACE FUNCTIONS (MESSAGES TO INTERFACE FUNCTIONS ARE DEFINED. MESSAGES FROM INTERFACE FUNCTIONS EXIST ACCORDING TO THE DESIGNER'S CHOICE)
6 REMOTE INTERFACE MESSAGES SENT BY INSTRUMENT FUNCTIONS WITHIN A CONTROLLER

fig. 7. Instrument designs can be conceptualized as being partitioned into two areas: instrument functions and interface functions. But this division does not necessarily imply two separate physical layouts within the instrument.

ham radio

the
weekender

noise cancellation circuit

As an LF enthusiast, I've often encountered QRN and QRM on the 160 to 190 kHz (1750 meters) band. In years past, the situation was manageable, but in recent years, it seems that conditions have grown steadily worse. A plague of EMI in my area renders weak signal reception impossible, and the source of much of the QRN, at least, has been increasing use of electric light dimmer switches.

In trying to alleviate the interference, I did some experimenting with synchronous line blankers. I also used the noise blanker in my Yaesu FT-101E with my LF converter. Nothing worked.

Recalling previous experience with direction finders, I decided to approach the problem in a different way, using a loop antenna, a sense antenna, and some sort of an RF phasing scheme. But my ferrite rods lacked either the length or permeability necessary to make a practical 1750 meter loop, so I tried the two-antenna scheme shown in **fig. 1**. I was pleasantly surprised; at the first attempt, I could reduce a 20 dB over S9 QRN to S7.

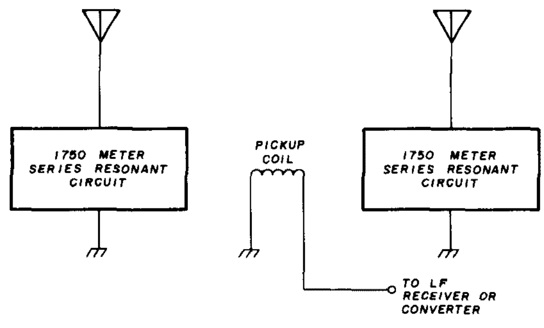


fig. 1. Two series-resonant circuits with a pickup coil gave surprising results.

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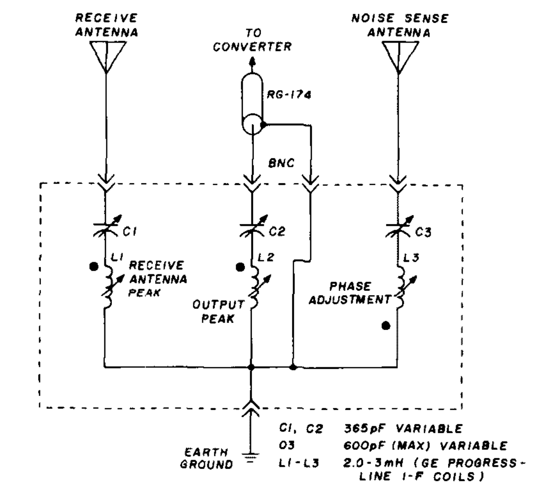


fig. 2. The 160-200 kHz RF phase cancelling circuit schematic. C1-C3 and L1-L3 can be increased or decreased in value to suit your frequency of interest. All coils are spaced 0.5 inch (12.7 mm) apart.

With a little further experimenting, I came up with a simple RF phase-cancelling circuit (**fig. 2**) that outperformed my Yaesu FT-101E noise blanker.

simple circuit layout

Three coils obtained from old General Electric two-way radio progress line IF transformers are mounted side by side in an enclosed metal minibox. The coils are approximately 0.5 inch (12.7 mm) apart; L1 and L3 are out of phase with each other. C1, C2, and C3 can be variable capacitors for wider frequency coverage, but for the 1750 meter band, I obtain adequate coverage by slug tuning the mutually coupled coils. Adjusting around the resonant frequency of both tuned circuits at the frequency of interest will result in phase cancelling of the unwanted noise with very little attenuation of the received signal.

cancelling noise

A careful adjustment can reduce 20 over 9 light dimmer noise to a noise level of S3 or less. Connect your "noise" antenna to the noise antenna input, then connect your LF receiver or LF converter to the output of the noise canceller. Leave the receive antenna off. Now tune the "phase" or noise antenna coil, L3, for maximum noise at the frequency of interest. Disconnect the noise antenna and connect the receive antenna. Peak the receive antenna coil, L1, for maximum noise at the frequency of interest. Connect both receive and noise antennas and adjust both coils carefully till your noise is nulled out. Repeak L2 for maximum signal.

I've had exceptional results. I've nulled out 20 over 9 light dimmer noise and managed to receive very weak LF experimental beacons, one from 67 miles away and another from 90 miles away. Even with my noise blanker on, I was never able to receive either of these before.

In some instances the noise blanker and noise canceller can be used together to cancel noise. I use a 260 foot wire for my receive antenna and a 50 foot inverted L configuration for my noise antenna. For best results, you may have to do some experimenting with your noise antenna configuration; it should be fairly long (at least 40 feet), with a good portion of its length running vertically.

The light dimmer in my house was the only noise source that the noise canceller couldn't phase out, but my easy access to the dimmer switch cured that problem. I simply replaced it with a conventional switch. Power line noise and light dimmer noise are no longer a problem to me, and 1750 meters is now fun to operate again.

For those interested in 1750 meters *The Lowfer Letter* is available free (send SASE) from Vincent J. Pinto, 2 Fairview Terrace, Suffern, New York 10901 — Editor

ham radio

speech synthesis for repeaters

Today's synthesizers are smaller, faster, smarter, cheaper

In the late 1960's a machine containing 60 endless loop tape cartridges was required for a human voice announcement of time. The device measured three feet high by nineteen inches wide. Today the same function is accomplished with a unit the size of a wrist watch, with all the electronics manufactured on a single wafer.

In the mid 1970's, the electronic industry began introducing electronic voice generating devices known as voice synthesizers. The initial versions, which the home experimenter could afford, fit on a 10 × 7-inch PC board and contained a limited vocabulary. But unlike its mammoth predecessor, it could be easily controlled by a computer or digital controller.

In the early 1980's, chip technology had advanced and the 10 × 7 inch module was reduced to a single device. Several manufacturers are now producing these devices and their use is commonplace in repeaters as well as the phone company. Amateur Radio manufacturers have started using these devices. The Kenwood TW-4000A 2 meter and 440 MHz transceiver features a voice option in both English and Japanese. Two Amateur repeater-controller manufacturers, Micro Security and Advanced Computer Controls (ACC)* also provide speech synthesis features.

If you dial Telephone Information for a phone number, you first speak with a human assistant. After taking your request, the assistant departs and your requested phone number is given by an electronic voice unit. Some expensive voice generators sound so much like human speech that only a musically-tuned ear can detect any difference. However, some less expensive units sound very good and are adequate for use in repeaters. One such unit is the National Semiconductor's "DT-1050 Digitalker."

*Advanced Computer Controls (ACC), 10816 Northridge Square, Cupertino, California 95014.

DT-1050 voice synthesizer

The DT-1050 contains a voice control processor and two 8K Byte ROMs. The control processor is a 40-pin IC which controls voice generation as instructed by the commands in the ROMs. The ROMs contain 143 characters; an outside controller can instruct the control processor to pronounce these characters via an 8-bit input latch. Each character is assigned an 8-bit combination that forces the processor to fetch the character data from ROM and execute the speaking of the chosen character. If a new character command is given, the processor will stop what it is doing and start with the new character. This feature allows forming words not in the existing vocabulary by using combinations of characters. An example is the word "repeater" which is very close to "re meter" pushed together. (I've even produced a reasonable "Cincinnati, Ohio" using *Cent, C, N, At, E, O, High*, and *O* pushed together.)

Table 1 provides a list of the DT-1050 master word list along with the character 8-bit data word commands. The first character, "This is Digitalker," is spoken in a "female" voice, but the remainder is spoken in a "male" voice. Note that the one through twenty character commands are 1 through 20 decimal. To force the character sixteen, the controller simply sends 16 decimal or 10 hexdecimal. Also, commands above 142 decimal or 8F hexdecimal are invalid. These commands will return garbage.

To control the DT-1050, one first places the character command on the 8-bit latched inputs SW1 (LSB) through SW8, and produces a low to high TTL signal on the "write" line (pin 4 of the processor). The processor then latches the 8 bits and starts the character speech sequence. It continues until finished or until a new command is entered, whichever comes first. When the processor is busy it produces a TTL low on the introduction

By Ron Wright, N9EE, Micro Security, 9307 Meadow Lane, Greenfield, Indiana 46140

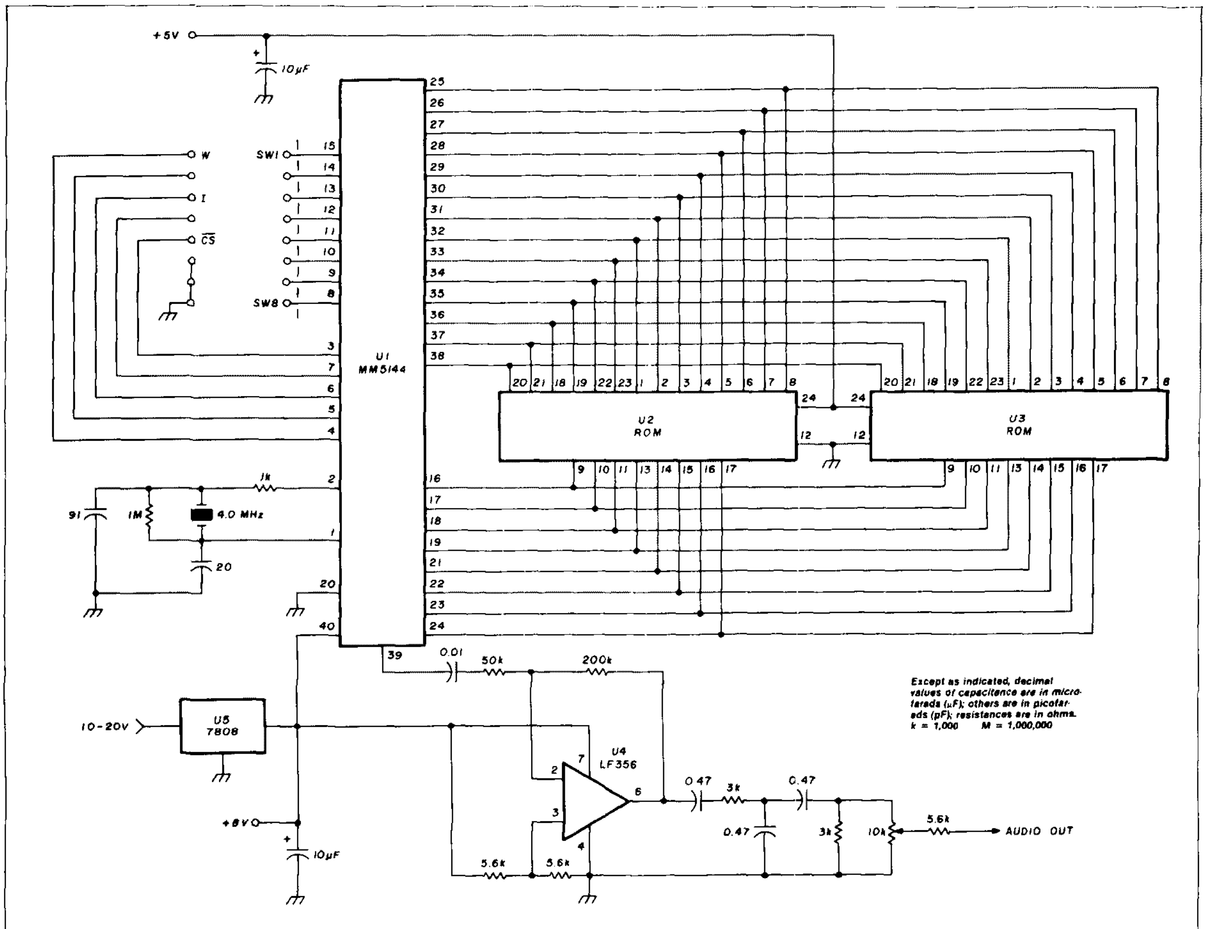


fig. 1. Voice generator DT-1050, audio filter and amplifier schematic.

(pin 6). When the character is complete, this line will go high. The external controller can use this line for "handshaking" with the voice processor.

The control processor can be tied directly to a microprocessor data bus with the Chip Select (pin 3) enabling the input latch. However, in the case of a hardware controller, this pin can be tied low. This pin controls only the input latch, enabling it when a command is to be entered. While the processor is speaking, the CS may be high or low.

the voice generator

Fig. 1 is a schematic of the DT-1050 and the needed audio filter and amplifier. The amplifier drives a 1 kilohm load. The voice processor controller operates on 7 to 11 volts and the ROMs require 5 volts. The 4 MHz crystal is the master clock and determines the speed at which the characters are pronounced. It is recommended that a 4 MHz crystal be used rather than the easy-to-find 3.58; there is a definite difference.

Fig. 2 is a schematic of a hardware controller for forming a list of phrases. The 2716 EPROM contains the character sequence. By programming the desired characters in the 2716 with each set in 16-byte blocks, one can produce and select desired character sequences. This controller uses only eight sets of the 16-byte blocks. The block addresses, which must be programmed with the desired voice sequence corresponding to the message input, are given in the inset table. If one wishes to expand the number of messages, a more detailed 2716 address control is required.

The 74374 latches the input command when the start line goes from low to high. This also presets the 7474 latch, which allows the 74123 to run clocking through the 2716 selected addresses, forcing the desired table to be outputted to the voice control processor. To start a voice sequence, first ground the desired 74374 "message input" line (M1 through M7). Then ground the "start" line, making sure it returns high prior to the message finish. The grounding of the "start" line and leaving the message lines high forces the table stored in 7F0 through 7FF to be sent. When a new byte is sent to the

table 1. DT-1050 master word list.

Word	8-Bit Binary Address SW8 SW1		8-Bit Binary Address SW8 SW1		SW8 SW1
THIS IS DIGITALKER	00000000	Q	00110000	IS	01100000
ONE	00000001	R	00110001	IT	01100001
TWO	00000010	S	00110010	KILO	01100010
THREE	00000011	T	00110011	LEFT	01100011
FOUR	00000100	U	W0110100	LESS	01100100
FIVE	00000101	V	00110101	LESSER	01100101
SIX	00000110	W	00110110	LIMIT	01100110
SEVEN	00000111	X	00110111	LOW	01100111
EIGHT	00001000	Y	00111000	LOWER	01101000
NINE	00001001	Z	00111001	MARK	01101001
TEN	00001010	AGAIN	00111010	METER	01101010
ELEVEN	00001011	AMPERE	00111011	MILE	01101011
TWELVE	00001100	AND	00111100	MILLI	01101100
THIRTEEN	00001101	AT	00111101	MINUS	01101101
FOURTEEN	00001110	CANCEL	00111110	MINUTE	01101110
FIFTEEN	00001111	CASE	00111111	NEAR	01101111
SIXTEEN	00010000	CENT	01000000	NUMBER	01110000
SEVENTEEN	00010001	400HERTZ TONE	01000001	OF	01110001
EIGHTEEN	00010010	80HERTZ TONE	01000010	OFF	01110010
NINETEEN	00010011	20MS SILENCE	01000011	ON	01110011
TWENTY	00010100	40MS SILENCE	01000100	OUT	01110100
THIRTY	00010101	80MS SILENCE	01000101	OVER	01110101
FORTY	00010110	160MS SILENCE	01000110	PARENTHESIS	01110110
FIFTY	00010111	320MS SILENCE	01000111	PERCENT	01110111
SIXTY	00011000	CENTI	01001000	PLEASE	01111000
SEVENTY	00011001	CHECK	01001001	PLUS	01111001
EIGHTY	00011010	COMMA	01001010	POINT	01111010
NINETY	00011011	CONTROL	01001011	POUND	01111011
HUNDRED	00011100	DANGER	01001100	PULSES	01111100
THOUSAND	00011101	DEGREE	01001101	RATE	01111101
MILLION	00011110	DOLLAR	01001110	RE	01111110
ZERO	00011111	DOWN	01001111	READY	01111111
A	00100000	EQUAL	01010000	RIGHT	10000000
B	00100001	ERROR	01010001	SS (Note 1)	10000001
C	00100010	FEET	01010010	SECOND	10000010
D	00100011	FLOW	01010011	SET	10000011
E	00100100	FUEL	01010100	SPACE	10000100
F	00100101	GALLON	01010101	SPEED	10000101
G	00100110	GO	01010110	STAR	10000110
H	00100111	GRAM	01010111	START	10000111
I	00101000	GREAT	01011000	STOP	10001000
J	00101001	GREATER	01011001	THAN	10001001
K	00101010	HAVE	01011010	THE	10001010
L	00101011	HIGH	01011011	TIME	10001011
M	00101100	HIGHER	01011100	TRY	10001100
N	00101101	HOUR	01011101	UP	10001101
O	00101110	IN	01011110	VOLT	10001110
P	00101111	INCHES	01011111	WEIGHT (Note 2)	10001111

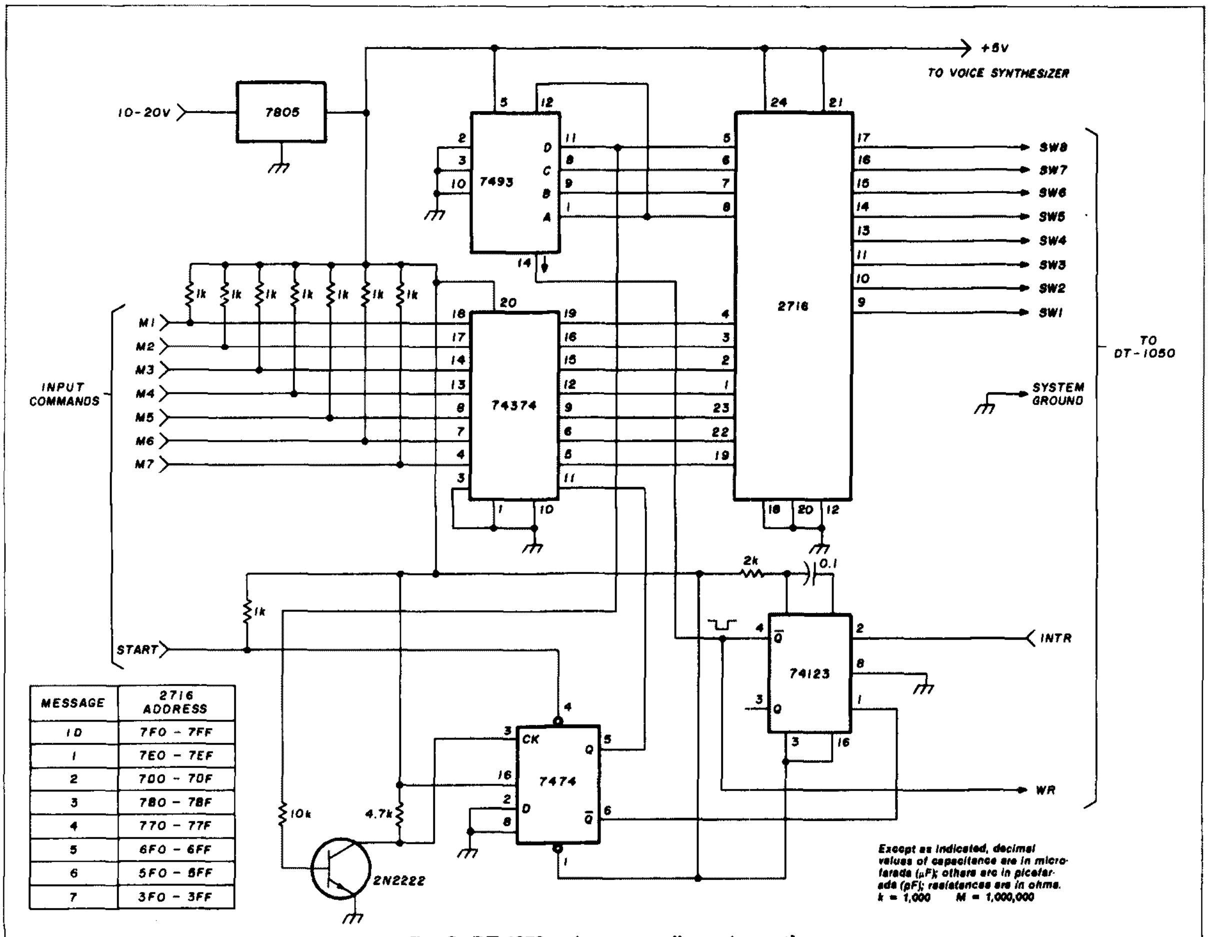
Note 1: "SS" makes any singular word plural

Note 2: Address 143 is the last legal address in this particular word list. Exceeding address 143 will produce pieces of unintelligible invalid speech data.

processor, the controller strobes the "write" line, beginning the character sequence. When the character is complete, the processor returns a low to high signal on the "intr" line, clocking the 74123, clocking the 7493, and selecting the next byte. This sequence continues until the 7493 reaches all ones (1111) and receives an additional strobe. This last strobe now forces a 0000 output and clocks the 7474, producing a stop command.

voice applications

Voice generator applications are many. The most obvious is the voice ID of a repeater. Other applications are "10 seconds time out" for repeater and autopatch time out warning; "you timed it out" on time out recovery; "2 meters on" for remote bases; and many more. If one has a microprocessor-based controller, one can add software with a minimal amount of hardware to produce



MESSAGE	2716 ADDRESS
ID	7F0 - 7FF
1	7E0 - 7EF
2	7D0 - 7DF
3	7B0 - 7BF
4	770 - 77F
5	6F0 - 6FF
6	5F0 - 5FF
7	3F0 - 3FF

fig. 2. DT-1050 voice controller schematic.

features such as a talking clock, which is touch-tone settable, or a touch-tone pad tester (the voice generator says "star" and "pound" as well as the remainder of the touch-tone digits). The list is limited only by our imaginations.

National is now marketing the DT-1057, a second set of ROMs without the voice processor, containing another set of characters. The voice synthesizer can be expanded, using all four ROMs. National is also trying to market software that will allow development of one's own character set, but rumor has it that the cost may be about $1000. National is also said to be willing to develop words for about $200 each, but the present set is certainly adequate for many repeater applications.

The DT-1050 can be purchased from any supply store that carries National Semiconductor products, though you may have to wait a few weeks for delivery.

conclusion

The rapid progress in this field has brought consumer products incorporating memory speech synthesis into nearly every American home. One can now buy a wrist watch with a built-in storage unit that will allow its owner to enter an 8-second message to be read back at a selected time; the message can remind the wearer of an appointment or simply remind him or her of some timely piece of information. Votan* is marketing a voice-recognition and generation unit for taking orders over the phone, in which the caller follows computer-generated instructions and the *computer* enters account codes and part numbers using the caller's voice. "You are the XYZ company," says the computer. "You have ordered fifteen items . . . shipment will be in four days." Next, an automatic stock-fetching unit collects the ordered items, boxes them up, weighs the package, places the correct amount of postage on it, attaches the mailing label, and sets it on the loading dock for shipment.

Speech synthesis has come a long way in a very short time. The technology is here and fully accessible for applications in Amateur Radio.

ham radio

*Votan, 4487 Technology Drive, Fremont, California 95438.

the effects and treatment of electric shock

Still checking for live circuits with your fingers? Read this

Even in this day of solid-state equipment, Amateurs need to be especially aware of the hazards of electric shock. We all know what causes shock, but not all Amateurs know just why an electric shock can be so dangerous.

Some people believe that an electric shock burns a victim to death. Yet, medical evidence suggests that currents strong enough to burn actually kill less often than do much lower voltages. Others think that the electric current "shorts out" a victim the same way that lightning may short out an electric circuit. While far from accurate, this latter belief is closer to what really happens than the former.

Our nervous system is a complex electrochemical system "masterminded" by the brain. Our various motor functions are controlled by minute electric signals sent along the complex network known as the nervous system; an electric current generally kills by overriding the controlling influence of this system.

The minute electrical impulses circulating in the nervous system lose control of body functions whenever they are overridden by an outside current. By applying small potentials to the brain, researchers have been able to induce the movement of limbs and also the stimulation of mental images. This electrical prodding has helped them learn much about the brain and its effects on human behavior. Not so helpful, however, are the uncontrolled currents that flow during an electric shock, for these currents can affect the brain's signals to vital body parts. Currents entering the heart and respiratory centers are especially dangerous; the key to surviving electric shock appears to lie not in how much current there is, but rather where it travels.

shock kills two ways

Death due to electric shock is generally caused by one of two effects: ventricular fibrillation or respiratory center paralysis. Ventricular fibrillation is irregular, arhythmic contraction of the muscles of the heart. The heart is a pump that forces blood throughout the body according to a rhythmic stimulus established in the right auricle or sinus node. This stimulus is a minute periodic electric current that flows to all sections of the heart and regulates the contractions of its muscles. If the system is upset — say by an outside current —

(continued on page 88)

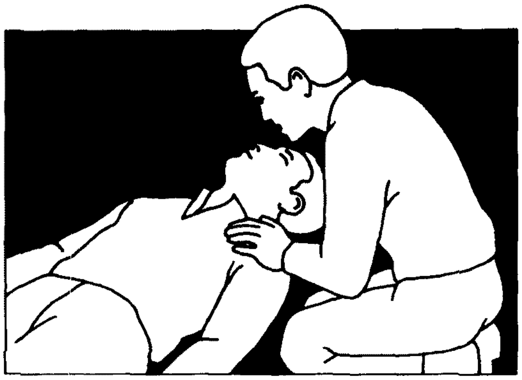
By Daniel Peters, Falcon Communications, P.O. Box 620625, Woodside, California 94062

YOU CAN LEARN CPR

When a person's heart and lungs stop functioning because of a heart attack, shock, drowning or other causes, it is possible to save that life by administering CPR, or cardiopulmonary resuscitation.

CPR provides artificial circulation and breathing for the victim. External cardiac compressions administered manually are alternated with mouth-to-mouth resuscitation in order to stimulate the natural functions of the heart and lungs.

This brochure contains an overview of CPR training and is not intended as a complete guide. Contact your local chapter of the AMERICAN RED CROSS for further information on how you can learn this life-saving procedure.

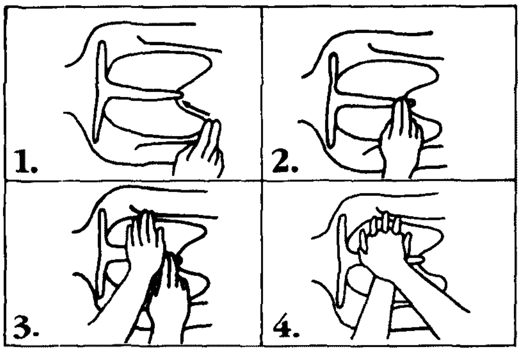


1. DETERMINE IF VICTIM IS UNCONSCIOUS

Tap or gently shake the victim's shoulder. Shout, "Are you O.K.?" If no response shout "HELP!" (Someone nearby may be able to assist.) Do the AIRWAY step next.

4. CHECK STEP

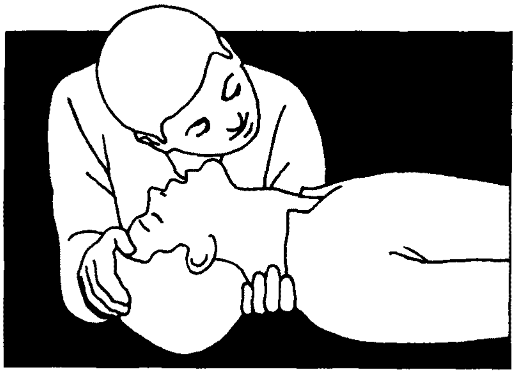
CHECK the pulse and breathing for at least 5 seconds but no more than 10. To do this, keep the head tipped with the hand on the forehead. Place the fingertips of your other hand on the adam's apple, slide your fingers into the groove at the side of the neck nearest you. If there is a pulse but no breathing give one breath every 5 seconds. If no pulse **or** breathing is present send someone for emergency assistance (dial 911 or operator) while locating proper hand position.
Begin Chest Compressions.



5. HAND POSITION FOR CHEST COMPRESSIONS

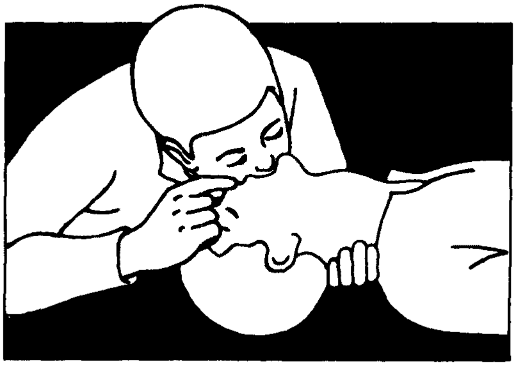
1. With your middle and index fingers find the lower edge of the victim's rib cage on the side nearest you.
2. Trace the edge of the ribs up to the notch where the ribs meet the breastbone.
3. Place the middle finger **on** the notch, the index finger next to it. Put the heel of the other hand on the breastbone next to the fingers.
4. Put your first hand on top of the hand on the breastbone. Keep the fingers off the chest.

fig. 1. CPR saves lives, but requires training and practice.



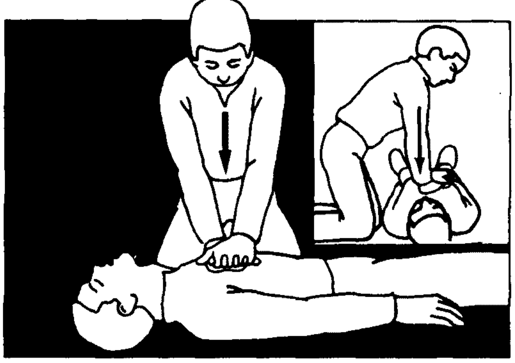
2. AIRWAY STEP

Place one hand on the forehead and push firmly backward. Place the other hand under the neck near the base of the skull and lift gently. Tip the head until the chin points straight up. This should open the airway. Place your ear near the victim's mouth and nose. LOOK at the chest for breathing movements, LISTEN for breaths and FEEL for breathing against your cheek. If no breathing occurs do the QUICK step next.



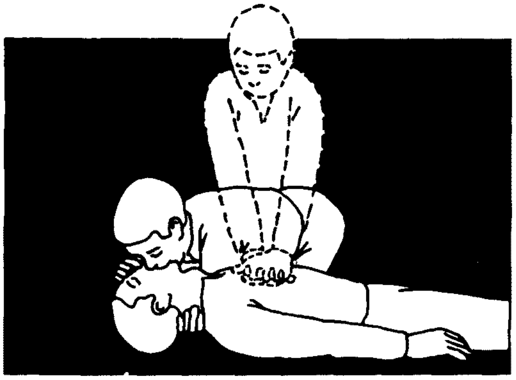
3. QUICK STEP

Give 4 QUICK full breaths, one on top of the other. To do this keep the head tipped and pinch the nose. Open your mouth wide and take a deep breath, making a good seal. Now, give the 4 breaths without waiting in between. Do the CHECK step next.



6. CHEST COMPRESSIONS

PUSH straight down without bending your elbows while maintaining proper hand position. Keep knees shoulder width apart. Shoulders should be directly over victim's breastbone. Keep hands along midline of body. Bend from the hip not the knees. Keep fingers off the chest. Push down about *1½ to 2 inches. Push smoothly.* Count, "1 and, 2 and, 3 and, etc.".



7. PUSH 15—BREATHE 2

Give 15 compressions at a rate of 80 per minute. Tip the head so the chin points up and give 2 quick full breaths. Continue to repeat 15 compressions followed by 2 breaths. Check the pulse and breathing after the first minute and every few minutes thereafter.
NOTE: **Do not practice** chest compressions on people as it could cause internal injuries.

THIS INFORMATION DOES NOT TAKE THE PLACE OF CPR TRAINING. CONTACT YOUR LOCAL RED CROSS CHAPTER ON HOW YOU CAN LEARN THIS LIFE-SAVING PROCEDURE.

the muscles of the heart may respond in an irregular fashion, rendering the organ useless as a pump. Without immediate trained medical attention, recovery from ventricular fibrillation is unusual.

Respiratory center paralysis is the second most lethal effect of electric shock. Normal breathing is controlled by the hindbrain; the stimulus travels from the brain through the nervous system to the lungs. An outside electric current can easily cause breathing to stop.

low voltage more dangerous?

Just how much current will cause death is difficult to determine. Much research has been done, but rather than list data that could be erroneously interpreted as a guideline of what constitutes safe levels, let's just say that under the wrong conditions, just a few thousandths of an ampere can do you in. But how does this equate to voltage?

I have tested the electrical resistance of my body under various conditions. One warm day, when I was perspiring freely, the resistance of my body (measured between two small pieces of pipe held in my hands) was low enough that a voltage as low as 25 volts *would have caused a current generally* considered lethal to a healthy adult. (Back in the days of 32-volt farm lighting systems, death from these systems was not unknown.)

At times shock from 1000 volts or more may be less dangerous than shock from lower voltages. The reason for this is that the high voltage, and attendant higher currents, cause all muscles — including those of the heart — to contract suddenly and violently. Sometimes the heart muscles may contract to such an extent that fibrillation can't occur; in such cases the heart may resume normal action if the current is stopped within three or four minutes. One report indicated a recovery *rate of 62 percent among persons* who were "knocked out" by voltages above 1000. The corresponding recovery rate at much lower voltages was only 39 percent.

People say a charged conductor "holds" its victim. It does this by contracting and paralyzing muscles. But in some cases muscles contract with enough force to "throw" the victim; this, of course, may cause secondary injuries if the victims's body strikes an obstruction as it falls.

watch out for that left hand

Current paths are extremely important in determining the *effects of* electric shock. Any route involving the heart or brain is particularly dangerous; leg-to-leg paths are considered less threatening.

In a study of cases involving fatal shock at voltages below 250, 90 percent of the victims had burns on their left hands. This suggests that shocks received through the left hand may kill more often than those through the right hand. (This statistic is even more noteworthy when you consider the predominance of right handed people in the general population.) So if you follow the rule of keeping one hand in your pocket while working on live circuits, make it your left hand. Better yet, *don't work on live circuits — ever.*

delayed results

If you receive a shock and *suffer no* apparent injury, your troubles are not necessarily over. Electric shock often damages delicate nerve cells in the spinal cord, which can cause a "wasting away" of muscle in one or more limbs. This is a slow, progressive, and intractable disturbance whose effects may not appear for weeks or even months after experiencing the shock. Other delayed effects may include insanity, personality changes, amnesia, mental inertia, blood-vessel diseases, cataracts, nerve disturbances, destruction of pancreatic tissue, and disruptions of the heart's conduction system.

how to help

If you are present when someone is rendered unconscious by an electric shock, the first thing to do is to ***stop the current flow***. If possible, do so by turning off the power at the nearest switch. If you can't do that, or if getting to the switch will take too long, the next best thing is to ***separate the victim from the source of current.*** This can be dangerous. Use a wooden board or other nonconducting object. As soon as you can touch the victim safely, begin cardiopulminary resuscitation (CPR) — but only if you've been trained in the technique. *Anyone who works near live circuits owes it to his or her colleagues to take a course in CPR, master the technique, and make sure the skills are practiced at regular intervals, or as needed. (See **fig. 1.**)*

Speed is essential. Any delay at all greatly reduces the chances of recovery. Of some 600 cases studied, over 70 percent of those receiving treatment within three minutes recovered. Just one more minute dropped the percentage of recovery to 58 percent. If there is no heart or respiratory action and treatment is delayed by five minutes, death is inevitable.

If you are alone, do not take the time to go for help. Start treatment immediately. If the victim can be saved, you are the one who will do it. ***Don't stop treatment — not even if the victim appears to be dead;*** in one case, eight hours passed before a victim responded. Only a physician can judge whether a victim is really dead.

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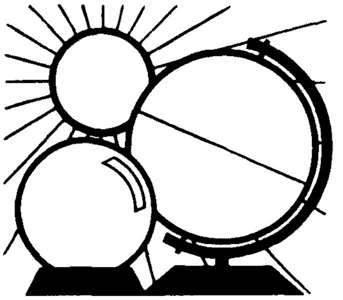

DX FORECASTER

Garth Stonehocker, K0RYW

equinox DX

The ionospheric structure that affects radio propagation changes rapidly during the two months of equinox (March and April). A second equinoctial ionospheric effect is related to the alignment of the equatorial planes of both the earth and the sun. This alignment allows more direct entry by solar wind particles to the earth's polar regions. The geomagnetic field modulated by these solar wind particles cause ions in the ionosphere to move. A signal ray traced through the ionosphere (see October, 1983, *DX Forecaster*) follows the ion's irregular path. The ray alternately focuses and defocuses its energy at the receive location. This is one cause of QSB, signal strength variability of a relatively short time (seconds to period minutes). Expect more intense and greater numbers of these geomagnetic-ionospheric disturbances this time of year.

Gray line DX, a very efficient type of twilight propagation for QRP work or at any power level, is aligned across the pole at this season. Look for this mode of propagation to occur within half an hour on either side of sunrise and sunset. (A beam bearing visualization aid, timing chart, and propagation descriptions can be found in *The Shortwave Propagation Handbook* by T. Cohen and G. Jacobs. This excellent book is available from Ham Radio's Bookstore, Greenville, New Hampshire 03048, for $9.95 postpaid.)

spring thunderstorm QRN

Winter has been very quiet; now noise — QRN — is here again. March and April are months in which the weather often consists of a series of spring storms bringing rain to much of our country. These storms are usually fronts of warm and cold air that generate the year's first major thunderstorms. These thunderstorms produce noise (static) that reduces the signal-to-noise ratio of received signals, worsening readability.

The cumulative effect of thunderstorm static worldwide is the main cause of high noise levels on lower frequency HF bands, mainly in summertime. However, as a specific storm front approaches, a significant increase in the noise level is heard. One first notices this increase in noise at a one-hop distance away, about 600 to 1200 miles (960 to 1920 km) when the storm front is about one day west of your location. Next, the noise level usually decreases as the storm moves closer. When the storm is within a ground-wave's distance (50 to 60 miles or 80 to 96 km), the noise level becomes more intense. Individual discharges can be heard. As the storm draws nearer, its sounds become part of the "local noise"; as it moves away, a similar decrease in noise occurs. An increase is heard again approximately one day later as the front reaches the one hop distance away. (You can check this out for your location by correlating information on storm movement given on the local television weather report with your operating and listening experience.) In looking for the rare DX, you may want to save time by tracking storms in order to determine the most favorable operating conditions.

last-minute forecast

The higher HF bands (10 to 30 meters) are expected to be excellent the first and last weeks of March. During the second and third weeks, the lower bands (30 through 160 meters) will be at their best. Disturbed geomagnetic and ionospheric conditions can be expected about March 7, 12, 17, 21, 26 and 31st. Spring equinox occurs on March 20th at 1024 UT. The moon is full on the 17th and at perigee on the 16th.

band-by-band summary

Ten, fifteen, and twenty meters will be open from morning to early evening almost every day, and to most areas of the world. The openings on the higher of the bands will be shorter and will occur more frequently near local noon. Transequatorial propagation on these bands will be more likely towards evening during conditions of high solar flux and a disturbed geomagnetic field. Noise effects are not too noticeable.

Thirty meters will be useful almost twenty-four hours a day. Daytime conditions will resemble those on 20 meters, although signal strength may decrease during midday on some days — days coinciding with high solar flux values. Nighttime use will be good except after days of very high MUF conditions. Generally the usable distance is expected to be greater than that achieved on 80 at night, but less than that on 20 meters during the day.

Forty, eighty, and one-sixty meters are the nighttime DXer's bands. The bands open just before sunset and last until the sun comes up on the path of interest. Except for daytime short-skip signal strengths, high solar flux values don't affect these bands much. Geomagnetic disturbances, however, which will be more evident near the equinox, cause signal attenuation and fading on polar paths. Noise will be sporadic and very noticeable on these lower frequency bands.

ham radio

MARCH

WESTERN USA

GMT	PST	N Asia Far East	NE Europe	E S. Africa	SE S. America	S Antarctica	SW New Zealand	W Oceania Australia	NW Japan
0000	4:00	20	20	15	10	15	10	10	20
0100	5:00	20	20	15	10	15	10	10	20*
0200	6:00	20	20	15	10	15	10	10	15
0300	7:00	20	20	15	10	15	10	10	15
0400	8:00	20	20	20	10	15	10	10	15
0500	9:00	20	20	20	10	15	10	10	15
0600	10:00	20*	30	20	10	20	10	10	20
0700	11:00	20*	*30*	20	10	20	10	15	20
0800	12:00	*20*	*30*	20	15	20	15	15	20
0900	1:00	*20*	*30*	*20*	15	20	15	15	20
1000	2:00	*20*	*30*	*20*	*15*	*20*	15	20	*30*
1100	3:00	*20*	*30*	*20*	*20*	*30**	*20*	20	*30*
1200	4:00	*30*	*30*	*20*	*20*	*30**	*20*	20	*30*
1300	5:00	*30*	*30*	15	*20*	*20*	*20*	*20*	*30*
1400	6:00	*30*	20	15	20	*20*	*20*	*20*	*30*
1500	7:00	30	20	15	20	*20*	*20*	*20*	*30*
1600	8:00	30	20	15	20	20	20	*20*	*30*
1700	9:00	30	20	10	20	15	15	20	*30*
1800	10:00	40*	20	10	15	15	15	15	20
1900	11:00	30	20	10	15	15	15	15	20
2000	12:00	30	20	10	15	15	10	15	20
2100	1:00	20	20	10	10	15	10	10	20
2200	2:00	20	20*	10	10	15	10	10	20
2300	3:00	20	20*	15	10	15	10	10	20

MID USA

GMT	MST	N Asia Far East	NE Europe	E S. Africa	SE S. America	S Antarctica	SW New Zealand	W Oceania Australia	NW Japan	CST
0000	5:00	30	20	15	15	15	10	10	20	6:00
0100	6:00	30	20	15	15	15	10	10	20	7:00
0200	7:00	30	20	15	15	15	10	10	20*	8:00
0300	8:00	30	20	15	15	15	10	10	20*	9:00
0400	9:00	30	20	20	20*	15	10	10	20*	10:00
0500	10:00	30	20	20	20	15	10	10	20	11:00
0600	11:00	*30*	30	20	20	20	10	15	20	12:00
0700	12:00	*30*	*30*	20	20	20	15	15	20	1:00
0800	1:00	*30*	*30*	20	20	20	15	15	20	2:00
0900	2:00	*20*	*30*	*20*	20	20	15	20	20	3:00
1000	3:00	*20*	*30*	*20*	*20**	*20*	*20*	20	*30*	4:00
1100	4:00	*20*	*30*	*20*	*15*	*20*	*20*	*20*	*30*	5:00
1200	5:00	*20*	*30*	*20*	*15*	*20*	*20*	*20*	*30*	6:00
1300	6:00	*20*	*30*	15	*15*	*20*	*20*	*20*	*30*	7:00
1400	7:00	20	20	15	10	*20*	*20*	*20*	*30*	8:00
1500	8:00	20	20	15	10	*20*	*20*	*20*	*30*	9:00
1600	9:00	20	20	10	10	20	20	*20*	*30*	10:00
1700	10:00	20	20	10	10	15	15	20	*30*	11:00
1800	11:00	20	20	10	10	15	15	15	20	12:00
1900	12:00	20	20	10	10	15	15	15	20	1:00
2000	1:00	20	20	10	10	15	10	15	20	2:00
2100	2:00	30	20	10	10	15	10	10	20	3:00
2200	3:00	30	20	10	15	15	10	10	20	4:00
2300	4:00	30	20	15	15	15	10	10	20	5:00

EASTERN USA

GMT	EST	N Asia Far East	NE Europe	E S. Africa	SE Caribbean S. America	S Antarctica	SW New Zealand	W Oceania Australia	NW Japan
0000	7:00	30	20	15	15*	10	10	10	20
0100	8:00	30	20	15	15	15	10	10	20
0200	9:00	30	20	15	15	15	10	10	20*
0300	10:00	30	20	15	15	15	10	10	20*
0400	11:00	30	20	20	20*	15	10	10	20*
0500	12:00	30	30	20	20	15	10	15	20
0600	1:00	*30*	30	20	20	20	15	15	20
0700	2:00	*30*	*30*	20	20	20	15	15	20
0800	3:00	*30*	*30*	20	20	20	15	*20*	20
0900	4:00	*30*	*40**	*20*	*20*	20	*20*	*20*	*30*
1000	5:00	*20*	*40**	*20*	*15*	*20*	*20*	*20*	*30*
1100	6:00	*20*	*30*	*20*	*15*	*20*	*20*	*20*	*30*
1200	7:00	*20*	*30*	*15*	*15*	*20*	*20*	*20*	*30*
1300	8:00	*20*	20	*15*	*15*	*20*	*20*	*20*	*30*
1400	9:00	20	20	15*	10	*20*	*20*	*20*	*30*
1500	10:00	20	20	10	10	20	*20*	*20*	*30*
1600	11:00	20	20	10	10	20*	20	*20*	*30*
1700	12:00	20	20	10	10	15	15	20	*30*
1800	1:00	20*	20	10	10	15	15	15	20
1900	2:00	20	20	10	10	15	15	15	20
2000	3:00	20	20	10	10	15	10	15	20
2100	4:00	30	20	10	10	15	10	10	20
2200	5:00	30	20	15	15	15	10	10	20
2300	6:00	30	20	15	15	15	10	10	20

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during "normal" hours.
*Look at next higher band for possible openings.

flea market

RATES Noncommercial ads 10¢ per word; commercial ads 60¢ per word **both payable in advance**. No cash discounts or agency commissions allowed.

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VIC 20 OWNERS. Let "Ham List" program jog your memory during a QSO. Instant search and find. Cassette, detailed operating instructions $8.95. WA2RDA, 415 Elm St., Fayetteville, NY 13066.

COUNSELORS: Connecticut brother-sister camp. Completely equipped with ham radio station. Program includes electronics, kit building, code and communications. June 25-August 22. Send resume: Lloyd Albin (N2DMQ) Ken-Mont and Ken-Wood Camps, 2 Spencer Place, Scarsdale, NY 10583.

FASTRAK™ 2004 Tone Decoder uses NE-567 PLL to detect single frequency tones for NOAA alerts or pilot presence; 3.6 × 1.5 inch pcb and manual $9.95. Proham Electronics Inc., 34620 Lakeland Blvd., Eastlake, OH 44094.

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Middlesex HA3 6HS
England

Holland Radio
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Republic of South Africa

NEED TO CONTACT James Navarchi concerning Yaesu gear. C.T. Huth, 146 Schonhardt St., Tiffin, OH 44883.

FOREIGN PAPER MONEY wanted for my hobby. Old and new. Will accept free or will buy or trade. Buddy Hincke, 1854 East Bay Drive, North Bend, Oregon 97459. WA6LFJ.

SUPER CQWW AND ARRL CONTEST PROGRAMS. TRS-80 Model I, III, (IV in III mode). Completely machine language. Automatic identification of country and zone (CQ) from call letters. Dupe speed 12000 + contacts per second. Screen displays *zones still needed (CQ), total points, zones, countries, etc.* Automatic CW generator with 2 buffers. Log print program prepares logs and dupe sheets. Log preparation program for hand logs. Similar features to above. QSL label program for both. CQ WPX now being written. FREE fact sheet and sample printouts. K4SB, 3496 Velma Drive, Powder Springs, GA 30073.

FOR SALE: New Cushcraft R3 halfwave vertical $215. Tom, WA1RTD, 21 Bayberry, Acton, MA 01720. (617) 263-2382.

120' GUYED TOWER. Extremely strong $700. 20' sections $150. Tim Colbert, 13609 Colony, Burton, Ohio 44021.

9.0 MHz SSB CRYSTAL FILTERS, 6 pole, 2.2 kHz bandwidth, 1.85 shape factor 6 to 60 dB. New, with hardware, specifications; $17.50 postpaid. 4CX250B chimneys, Johnson #124-0111-001, new, boxed; $5.00 postpaid, two for $9.00. Dentron scout C.A.P. *transceiver, new, $300.00 postpaid.* Mosley CM-1 receiver, 80-10, VGC, $60.00 postpaid. Hammarlund SP600JX-17, G.C. $140.00. W.E. Delage, PO Box 231, Kent, Ohio 44240.

VLF-LF preamps, coupler, Loran-C boards. SASE. Burhans Electronics, 161 Grosvenor St., Athens, Ohio 45701.

OSL CARDS: $17.50/500, ppd. Free catalog. Bowman Printing, 743 Harvard, St. Louis, MO 63130.

ELECTRON TUBES: Receiving, transmitting, microwave . . . all types available. Large stock. Next day delivery most cases. Daily Electronics, 14126 Willow Lane, Westminster, CA 92683. (714) 894-1368.

WANTED: 2 new 813 sockets and 15 to 25 MFD, 4000VDC oil filled capacitor. Bill Blake, W5SCM, Star 222, Columbus, MS 39701.

RUBBER STAMPS: 3 lines $4.50 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

RECEIVERS — Motorola WWV $35. Hammarlund HQ-100A $85. National NC-300 $95. R.F. Signal Gen. 80 kc-60 mc $30. H.P. Audio Signal Gen. $30. K6KZT, 2255 Alexander, Los Osos, CA 93402.

CHASSIS and cabinet kits. SASE K3IWK.

GD5UG AND XYL offer the use of their home, car, and station to a Ham living near to the U.S. West Coast in exchange for the use of a motor home for approximately three months in late mid 1984. OK in Callbook.

CABLE CONVERTERS, decoders. Catalog $1 refundable. APS, POB 263 HR, Newport, RI 02840.

WANTED: Cash paid for used speed radar equipment. Write or call: Brian R. Esterman, PO Box 8141, Northfield, Illinois 60093. (312) 251-8901.

RECONDITIONED TEST EQUIPMENT $1.00 for catalog. Walter, 2697 Nickel, San Pablo, CA 94806.

LINEAR: Gates HFL2500 with manual. Best offer. K3BN, 4946 Manor Lane, Ellicott City, MD 21043. (310) 995-1252.

RTTY-EXCLUSIVELY for the Amateur Teleprinter. One year $7.00. Beginners RTTY Handbook $8.00 includes journal index. P.O. Box RY, Cardiff, CA 92007.

DRAKE TR-7. Full set of filters, noise blanker, fans, AUX-7, PS-7 power supply, remote VFO, service manual, extended card set, and matching speaker. $1,000. Also MN-7 antenna tuner, $100; and SP-75 speech processor, $90. Reasonable offers considered. Lawrence Sires, 1229 Iowa Ave., Ridgecrest, CA 93555.

IMRA International Mission Radio Assn. helps missioners — equipment loaned; weekday net, 14.280 MHz, 2-3 PM Eastern. Br. Frey, 1 Pryer Manor Rd., Larchmont, NY 10538.

NATIONAL AACS ALUMNI NET meets every Friday at 2200 UTC on 14297 kcs, alternate 14287 kcs, USB. At 2230 UTC on 21397 kcs, alternate 21387 kcs USB. PACIFIC AACS ALUMNI NET, Mondays, Wednesdays, Fridays at 1800 UTC (10 AM PST) on 7218 kcs LSB. K6RG and W6ZF NCS.

"HAMS FOR CHRIST." Reach other Hams with a gospel tract sure to please. Clyde Stanfield, WA6HEG, 1570 N. Albright, Upland, CA 91786.

AMSAT is looking for copies of March, May, June, July 1916 QST. *If you have them, please contact* Wm. Lazzoro, N2CF, at AMSAT, 850 Sligo, Suite 601, Silver Springs, MD 20910.

QUICK-FIND Callsign log. Quickly know if and when you worked that call, and if you want to work it again! $2.00, full price, Quick-Find, 2725-H Sandicrest, Cantonment, Florida 32533.

TENNATEST — Antenna noise bridge — out-performs others, accurate, costs less, satisfaction guaranteed. Send stamp for details, W8URR, 1025 Wildwood Road, Quincy, MI 49082.

WANTED: Old RCA, Western Electric tubes. (713) 728-4343. Maury Corb, 11122 Atwell, Houston, Texas 77096.

WANTED: Early Hallicrafter "Skyriders" and "Super Skyriders" with silver panels, also "Skyrider Commercial", early transmitters such as HT-1, HT-2, HT-8, and other Hallicrafter gear, parts, accessories, manuals. Chuck Dachis, WD5EOG, The Hallicrafter Collector, 4500 Russell Drive, Austin, Texas 78745.

SELL: Kenwood Twins. E. Alline, NE5S, 773 Rosa, Metairie, LA 70005.

VERY In-ter-est-ing! Next 4 issues $2. Ham Trader "Yellow Sheets", POB356, Wheaton, IL 60189.

IBM-PC ASCII/BAUDOT/CW. SASE for details. E. Alline, NE5S, 773 Rosa, Metairie, LA 70005.

ANNIE's EASY. Analyze dipoles, slopers, verticals, inverted-vees and arrays; any orientation, position, phasing, weight or combination with Annie Antenna Analysis Software. Include REAL GROUND (conductivity, dielectric constant). Superb hi-res plotting. Annie's incredibly friendly and with 100% machine language, she's FAST! For Apple II + (48K or 64K) or IIe, DOS3.3, $49.95 + $2.00 postage, NY add sales tax. Include full name and call. S.A.S.E. for info. Commercial, library, etc., call for quote (315)622-3641. Sonnet Software, Dept. HR, 4397 Luna Course, Liverpool, NY 10388.

Coming Events

ACTIVITIES

"Places to go..."

COLORADO: The Grand Mesa Repeater Society's fifth annual Western Slope Hamfest, Saturday, March 31, 10 AM to 4 PM, Plumbers and Steamfitters Union Hall, 2384 Highway 6 and 50, Grand Junction. Free admission. Swap tables $5.00 each. Novice exams, tech sessions, auction and refreshments. Talk in on 146.82 and 449.20. For information or swap tables SASE to Larry Brooks, WB0ECV, 3185 Bunting Avenue, Grand Junction, CO 81504. (303) 434-5603.

ILLINOIS: LAMARSFEST 1984, Sunday, March 25, Lake County Fairgrounds, Grayslake, Rts. 45 and 120. Setup 6 AM. Public admitted 8 AM. Tickets $2.00 advance. $3.00 at door. 8 ft. tables $5.00 each — reservations encouraged. Talk in on 147.63-03 or 146.94 simplex. For information, tickets, reservations: SASE to LAMARS, PO Box 751, Libertyville, IL 60048.

ILLINOIS: The Sterling-Rock Falls Amateur Radio Society's 24th annual Hamfest, March 18 at the Sterling High School Fieldhouse, 1608 Fourth Avenue, Sterling. Commercial distributors, dealers and a large flea market. Doors open 7:30 AM. Tickets $2.00 advance, $3.00 at door. Flea market tables requiring electricity and all commercial tables $5.00. All others $3.00. Concession stand and accommodations for overnight self-contained campers available. For tickets, tables or information: Sue Peters, KA9GNR, PO Box 521, Sterling, IL 61081 or call (815) 625-9262. Talk in on W9MEP 146.25/85.

ILLINOIS: The 18th annual Rock River ARC Hamfest, Sunday, April 8, Lee County 4-H Center. Advance ticket donation $2.00. $3.00 at gate. Tables available for $5.00 (8 ft). Camping space available for nominal charge. Breakfast and lunch will be served and an auction for amateur related gear. Talk in on 37/97 repeater. Doors open 8 AM. For information, tickets, tables write or call Shirley Webb, KA9HGZ, 618 Orchard St., Dixon, IL 61021. (815) 284-3811.

INDIANA: The Morgan County Repeater Association Club's Martinsville Hamfest, March 11, Indiana State Fairgrounds Pavillion Building, Indianapolis. Admission $4.00 at the door. Tables: premium $30.00, flea market $8.00. Flea market space/no table $1.00. All tables advance reservations. Talk in on 147.21 and 146.52 simplex. For information and table reservations SASE before March 1 to Aileen Scales, KC9YA, 3142 Market Place, Bloomington, IN 47401.

INDIANA: The Putnam County Amateur Radio Club's second Amateur Radio and Electronics auction, April 7, Putnam County Fairgrounds, north of Greencastle. Doors open at 8 AM, auction at 10 AM. 5% commission on sales. Food available. Talk in on 147.93/.33. For information: John Underwood, K9IIB, RFD 1, Box 10, Fillmore, IN 46128.

MASSACHUSETTS: The Framingham ARA's annual Spring Flea Market, Sunday, April 1, Framingham Civic League Bldg., 214 Concord St. (Rt. 126) downtown Framingham. Doors open 10 AM, sellers setup 8:30. Admission $2. Tables $10 — pre-registration required. Radio equipment, computer gear, bargains galore! Talk in on 147.75/15 and 52. Contact Jon Weiner, K1VVC, 52 Overlook Dr., Framingham, MA 01701 (617) 877-7166

MASSACHUSETTS: The Wellesley Amateur Radio Society's annual Auction, Saturday, April 14, First Congregational Church of Wellesley Hills, 207 Washington Street, Wellesley. Doors open 10 AM and the auction begins at 11 AM. Talk in on 63-03, 04-64 and 52. Contact: Kevin P. Kelly, WA1YHV, 7 Lawnwood Place, Charlestown, MA 02129

MICHIGAN: The 23rd annual Michigan Crossroads Hamfest, sponsored by the Southern Michigan Amateur Radio Society and the Marshall High School Photo-Electronics Club, Saturday, March 24 from 8 AM to 3 PM, at the Marshall High School. Tickets $2.00 at the door; $1.50 advance. Table space 50¢ per ft., min. 4 ft., reserved til 8 AM. Snack bar and full food service. For reservations SASE to: SMARS, PO Box 934, Battle Creek, MI 49016 or call Wes Chaney, N8BDM (616) 979-3433. Talk in on 146.52 and 146.07/67.

MICHIGAN: The South Eastern Michigan Amateur Radio Association (S.E.M.A.R.A.) 26th annual Hamfest Swap and Shop, April 8, 8 AM to 3 PM, Grosse Pointe North High School, Vernier Rd. between Mack and Lakeshore. Advance admission $1.00, $2.00 at door. Talk in on S.E.M.A.R.A. repeater 147.75/15. For further information SASE to S.E.M.A.R.A., PO Box 646, St. Clair Shores, MI 48083, or call WD8OVL (313) 445-8651

MISSOURI STATE ARRL Convention, Kansas City, April 7-8, 1984. For information write or call PHD Amateur Radio Association, PO Box 11, Liberty, MO 64068-0011 (816) 781-7313.

NEW JERSEY: The Delaware Valley Radio Association's 12th annual Flea Market and Computer Show, Sunday, April 1, 8 AM to 4 PM, New Jersey National Guard 112th Field Artillery Armory, Eggerts Crossing Road, Lawrence Township, Trenton. Advance registration $2.50 or $3.00 at door. Indoor/outdoor flea market. Dealers and refreshments. Sellers bring own tables. Talk in on 146.52 and 146.07-.67. For tickets and space reservations SASE to KB2ZY, 140 Susan Drive, Trenton, NJ 08638.

NEW JERSEY: Ham Radio Flea Market sponsored by the Chestnut Ridge Radio Club, Saturday, March 24, Education Building, Saddle River Reformed Church, East Saddle River Road and Weiss Road, Upper Saddle River. Tables $10.00 for first, $5.00 each additional. Tailgating $5.00. Free admission. Food and soda available. Contact: Jack Meagher, W2EHD (201) 768-8360 or Roger Soderman, KW2U (201) 666-2430.

NORTH CAROLINA: The Raleigh Amateur Radio Society's 12th annual Hamfest and Flea Market, Sunday, April 15, Crabtree Valley Shopping Mall, intersection of US70 west and US 1&64. Starts 8 AM. Admission $4.00 at gate, no extra charge for tailgaters. Tables available for rent. There will be special interest meetings, CW and homebrew contests. Plenty of nearby hotels, restaurants, etc. Talk in on W4DW 146.04/146.64 and K4ITL 146.28/146.88. For more information: Pete Thacher, N4HQZ (919) 876-4073 or Jim Bradley, WA4ACO (919) 851-2437 6-8 PM weekdays or weekends or write RARS, PO 19127, Raleigh, NC 27619

OHIO: The Lake County Amateur Radio Association's fifth annual Lake County Hamfest and Computerfest, Sunday, March 25, Madison High School, Madison. Open for exhibitors at 5:30 AM and for the public 8 AM to 4 PM. Admission $3.00 advance and $3.50 at door. Tables $5.00 per 6 ft. and $6.50 per 8 ft. Check in and talk in on 147.81/.21. For information and reservations SASE to Lake County Hamfest Committee, PO Box 150, Mentor, Ohio 44061 (216) 953-9784

TEXAS: The San Antonio Area Radio Club's FIRST annual Swapfest and Bar-B-Q, April 7 at Comanche Park, 7 AM to 5 PM. Talk in on 147.36 MHz. For details: Melvin Anderson, 8932 Saddle Trail, San Antonio, Texas 78255

TEXAS: The Midland Amateur Radio Club's annual St. Patrick's Swapfest, Saturday, March 17 at 10 AM to 6 PM and Sunday, March 18, 8 AM to 2:30 PM, Midland County Exhibit Building east of Midland. Pre-registration $5, $6 at door. Tables $6 each. Refreshments available. Talk in on 16/76 and 33/93. For information or reservations contact Midland ARC, PO Box 4401, Midland, Texas 79704

WISCONSIN: The Madison Area Repeater Association (M.A.R.A.) will hold its 12th annual Swapfest, Sunday, April 8, Dane County Exposition Center Forum Bldg. in Madison. Doors open at 5 AM for commercial exhibitors, 8 AM for flea market sellers and 9 AM for general public. Equipment and components for hams, computer hobbyists and experimenters. Admission $2.50 advance and $3.00 at the door. Children twelve and under admitted free. Flea market tables $4.00 each in advance and $5.00 at the door. An all-you-can-eat pancake breakfast and a Bar-B-Q lunch will be available. Talk in on WB9AER/R, 146.16/.76. For reservations or information: M.A.R.A., PO Box 3403, Madison, WI 53704.

OPERATING EVENTS

"Things to do..."

MARCH 24 TO 26: B.A.R.T.G. Spring RTTY Contest, 0200 GMT Saturday until 0200 GMT Monday. Total contest period is 48 hours but not more than 30 hours of operation is permitted. Bands: 3.5, 7.0, 14.0, 21.0, and 28 MHz. Stations may not be contacted more than once on any one band but additional contacts may be made with the same station if a different band is used. Messages: Time GMT, RST and contact number. All logs must be received by May 31, 1984 to qualify. Summary and log sheets available from contest manager for two IRC's: Peter Adams, G6LZB, 464 Whippendell Road, Watford, Herts, England WD1 7PT

CONNECTICUT QSO PARTY sponsored by the Candlewood ARA from 2000Z March 31 until 0200Z April 1 with a rest period from 0500 to 1200Z. Suggested frequencies: Phono — 3927, 7250, 14,295, 21,370, 28,540. CW — 40 kHz from low end. Novice — 3725, 7125, 21.125, 28.125. Mail by April 30, 1984 to CARA, c/o R. Dillon, N2EFA, PO Box 954, Danbury, CT 06810.

NEW products

TC-1 plus ATV transmitter/downconverter

P.C. Electronics has upgraded its TC-1 420-450 MHz full color ATV unit with some new features plus many options that were once offered at an additional cost now standard, at no increase in price.

With more Amateurs using computers and VCRs on ATV, separate video and audio inputs were added to the existing camera and mic inputs. This allows front-panel switching back and forth between the camera and computer, or transmitting both the VCR audio along with voice over commenting using a microphone. It has made learning Basic over the air, as well as retransmitting space shuttle video and audio, *easy*.

Capability for external 13.8 VDC in addition to the built-in AC supply has been provided for those who want to go mobile or portable on battery power during field day, emergency services, CAP searches, parades, marathons, or other public service events.

The TC-1+ has the new TXA5-5 exciter/modulator which features two-frequency plug-in crystal switching with just the addition of a SPST switch. The built-in sync stretcher and high/low power switch capability enable superior stable color video if a higher power linear

amplifier, such as the Mirage 100 watt D1010N, is added later or run barefoot at its greater than 10 watt PEP RF output.

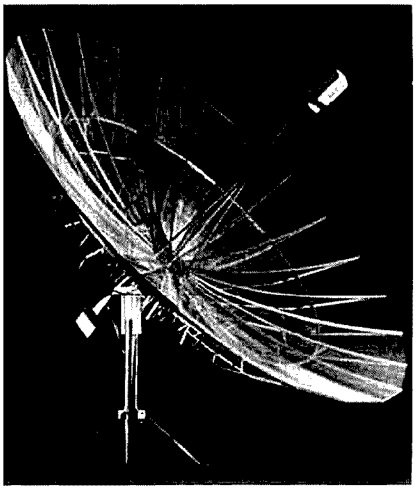
The 420-450 MHz tuneable downconverter has the low-noise NE64535 preamp stage to dig out the weak signals. It acts like a super hot UHF TV tuner, but covering only the 70 cm ham band, when connected to your TV set antenna input and set for channel 3 or 4. Both color video and sound live action Amateur television (ATV) are available on your TV set just as the broadcast stations provide. The standards are the same. A Technician class or higher Amateur Radio license is required for operation and purchase from P.C. Electronics.

More information and a complete catalog of ATV equipment, antennas, cameras, modules, and accessories are available from P.C. Electronics, 2522 South Paxson, Arcadia, California 91006.

Circle #301 on Reader Service Card.

parabolic antenna

The enlarged surface of the X-16 Parabolic Satellite Antenna by KLM Electronics, Inc., gives greater signal gain to compensate for weaker signal strength in locations on the fringe of satellite footprints — generally, areas outside of the continental U.S. Its modular aluminum construction permits assembly by Amateurs using ordinary tools. It is easily attached to its gold-anodized steel X-11 Polar-Trak mount with precision motor for accurate satellite tracking. Remote control is provided by a standard Polar-Trak with east-west pushbutton control or by KLM's programmable Memory Trak Dish Control Console which remembers fifty satellite locations.



The X-16 is available in "decorator" colors including forest green, desert tan, and brown, and can be ordered in other hues. Suggested retail for the X-16 and mount is $2,195; for an entire system, $4,795.

For details, contact KLM Electronics, Inc., 16890 Church Street, Morgan Hill, California 95037.

Circle #302 on Reader Service Card.

30-channel programmable scanner

Regency Electronics now offers a computer-controlled 30-channel programmable scanner loaded with advanced features for monitoring the action on more than 15,000 frequencies. Manufactured in the United States (Regency is the only American-made scanner), the Regency Model DX 3000 is available for $269.95 at participating Regency Electronics dealers.

The Regency DX 3000 covers six bands: low and high VHF (30-50 and 148-174 MHz), UHF (450-470 MHz), UHF "T" (470-512 MHz) and two FM ham bands (144-148 and 440-450 MHz); no crystals are required, and a CMOS memory with battery back-up saves frequencies for up to six months in the event of a power outage or if the scanner is stored.

Any selected frequency can be programmed into any selected channel with a few keys. The DX 3000 can search automatically (every 5 kHz on VHF, 12.5 kHz on UHF) for an active frequency; when it finds one, it pauses for four seconds to allow time for the operator to either enter it into memory or jot it down for reference.

The DX 3000 searches at the rate of 400 VHF frequencies (2 MHz) in about 34 seconds and 400 UHF frequencies (5 MHz) in about 30 seconds. Scan rate for the 30 programmed channels is about 15 channels per second. A channel lockout feature can exclude any selected channel(s) from being scanned; this keeps generally busy channels programmed into the scanner while preventing them from "locking in" on each scan. A selected *priority channel* is sampled every two seconds. If active, it automatically overrides any other signal. The scan delay feature holds the channel open for approximately two seconds at the end of a transmission to wait for any reply; if scan delay is not selected, scanning resumes in about six-tenths of a second.

Programming the DX 3000 is simplified by a series of plain-language messages that appear on its vacuum fluorescent digital display. These messages identify its current status and prompt appropriate actions. Operation is made easy with up-front controls for on-off/volume and squelch. Dual built-in power supplies permit plug-in AC operation, and 12 volt DC operation where not prohibited by law. UL listed and FCC certified (Part 15, Subpart C), the DX 3000 measures 10 1/3 × 3 1/3 × 7 inches.

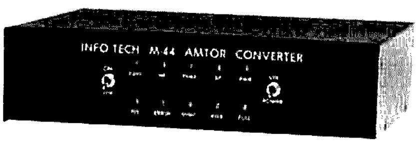
For additional information, contact Regency Electronics, Inc., 7707 Records Street, Indianapolis, Indiana 46226-9986.

Circle #303 on Reader Service Card.

AMTOR converter

The new Info-Tech M-44 AMTOR convertor allows most RTTY terminals to be used on the recently approved AMTOR RTTY mode. Interface to the terminal is via serial TTL or RS-232 levels, and either ASCII or Baudot terminals may be used.

The unit also freatures a built-in modulator and demodulator with pre-filter, full time ATC, and two transmit buffers. All control of the M-44 and transceiver are simple commands entered via the terminal keyboard.



Priced at $379.95, this converter is American designed and manufactured and will operate in the ARQ, FEC, and ARQ monitor modes.

For information, contact Digital Electronics Systems, 1633 Wisteria Court, Englewood, Florida 33533.

Circle #304 on Reader Service Card.

NEW products

IC-27A two-meter mobile

The ICOM IC-27A measures only 1-1/2 inches wide by 5-1/2 inches high and contains an internal speaker, making it easy to mount. Although the IC-27A is the most compact two-meter mobile unit on the market, no features have been sacrificed. Standard features include 25 watts of output power, 32 PL frequencies, ten full-function tunable memories, scanning of memories and the band, priority scan and a microphone which includes a 16-button touchtone pad for easy access to a repeater or dialing through to an autopatch. An optional speech synthesizer is also available to verbally announce the receiver frequency of the transceiver through the simple push of a button.

The IC-25A, measuring 2 inches wide by 5-1/2 inches high, will continue to be available for those individuals preferring a 25-watt two-meter mobile unit with larger operating knobs.

For more information, contact ICOM, 2112 116th Avenue, N.E., Bellvue, Washington 98004.

Circle #305 on Reader Service Card.

new triband beams

Two triband beams previously unavailable in the U.S. are now available from Palomar Engineers. Long a favorite of European DX'ers, the Model DX-33 (with three elements on 10, 15, and 20 meters) and DX-43 (with four elements) were designed for use with solid-state transceivers. The antennas feature low SWR, wide bandwidth, and particularly good gain and front-to-back ratio. Each trap is individually sweep-tested at the factory for uniform performance. Stainless steel "U" bolts are used throughout.

For more information, contact Palomar Engineers, 1924-F West Mission Road, Escondido, California 92025.

Circle #306 on Reader Service Card.

low-attenuation coax

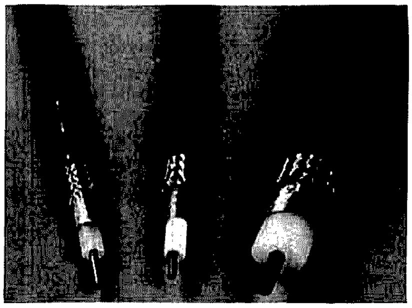
Belden Electronic Wire and Cable has available three 50-ohm low-attenuation, flexible coax cables (Belden 9913, 9914, 9915) for cellular radio, satellite communications, microwave and other two-way communications. Designed as flexible alternatives to semi-rigid cable to allow for ease of installation while maintaining similar electrical parameters, the cables will fit standard connectors.

Belden 9913, an RG-8/U type air dielectric coax, has an attenuation of 4.5 dB at 1 GHz, 11

NEW products

dB at 4 GHz, and 21 dB at 10 GHz. Nominal capacitance is 24 pF per foot. Overall diameter is 0.405 inch. Standard put-ups are 100, 250, 500, and 1000 feet. Representative price is $417.75 for 1000 feet.

Belden 9914, an RG-8/U type foam dielectric coax, has an attenuation of 1.6 dB at 100 MHz, 3.1 dB at 300 MHz, 4.1 dB at 500 MHz, 5 dB at 700 MHz, 6 dB at 1 GHz, 13 dB at 4 GHz, and 25 dB at 10 GHz. Nominal capacitance is 26 pF per foot. Overall diameter is 0.405 inch. Standard put-ups are 100, 250, 500, and 1000 feet. Representative price is $414.15 for 1000 feet.



Belden 9915, an RG-218/U type solid polyethylene insulated coax, has an attenuation of 0.83 dB at 100 MHz, 1.6 dB at 300 MHz, 2.4 dB at 500 MHz, 2.7 dB at 700 MHz, 3.5 dB at 1 GHz, and 10 dB at 4 GHz. Nominal capacitance is 30.8 pF per foot. Overall diameter is 0.870 inch. Standard put-ups are 250 and 500 feet. Representative price is $1,186.50 for 500 feet.

For additional information, contact Belden Electronic Wire and Cable, 2000 South Batavia Avenue, Geneva, Illinois 60134.

Circle #307 on Reader Service Card.

MR4 receiver

The new Micro Control Specialties MR4 receiver uses seven helical resonators in the front end and twelve poles of IF filtering to achieve a dynamic range greater than 100 dB. By using a first IF of 21.4 MHz and extensive shielding, it also rejects images by 120 dB. Other features of the MR4 include automatic slow/fast squelch, squelch hysteresis, and metering circuitry for signal strength, peak deviation, and discriminator.

The MR4 is intended for fixed frequency applications in the VHF and UHF bands, especially at multi-transmitter sites where RF interference is severe. The receiver is available in both modular and rack-mounted versions. The rack-mounted version includes full metering plus a local audio speaker.

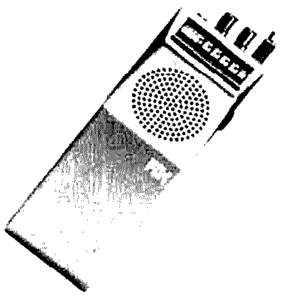
For more information, contact Micro Control Specialties, 23 Elm Park, Groveland, Massachusetts 01834.

Circle #308 on Reader Service Card.

NEW products

PLL-synthesized VHF FM receiver

FDK International Corporation has developed a PLL-synthesized VHF FM receiver, the RX-40, covering 141 MHz to 180 MHz bandrange or virtually all main VHF FM frequencies. Designed for use in Amateur, commercial, and marine band communication, it uses PLL-synthesized circuitry to provide accurate frequency selection of 15,600 channels between 141-180 MHz in 2.5 kHz steps. Its light weight (11 ounces; 315 g.) and miniature size (6-5/8 × 2-3/8 × 1-5/8 inches; 169 × 58 × 43 mm) allow maximum portability. Supplied with Ni-Cd battery pack, flex rubber antenna, AC



charger and earphone, the RX-40 features an adjustable squelch level to eliminate background noise on the AM mode and offers extremely low battery consumption providing continuous operation for ten hours. A BNC aerial connector, DC charger and shoulder case are also available as operational accessories.

For further information, contact FDK International Corporation, 10-2, Kaji-cho 2-chome, Chiyoda-ku, Tokyo 101, Japan.

Circle #309 on Reader Service Card.

antenna analysis software

"Annie" antenna analysis software, available for the Apple II+ and IIe with DOS 3.3, can calculate the patterns of nearly any wire antenna including dipoles, verticals, inverted vees, slopers and arrays of any length, orientation, position, power, phase or combination. "Annie" even includes the effects of real ground — conductivity and dielectric constant. It plots horizontal, vertical, or total gain in any direction; the plots may be drawn at any magnification or with any aspect ratio (for truly round circles). Any number of patterns may be drawn on each grid. The patterns can be drawn

with solid, dotted, dot-dashed or dashed lines. The plot can be printed at an FX-80 printer — the graphics dump source is included for modification for use of other printers.

In addition to horizontal, vertical, and total gain, "Annie" will calculate and tabulate polarization sense, axial ratio, tilt and phase. Any quantity may be printed in any column.

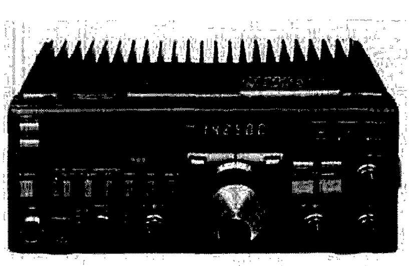
Written in assembly language for speed, the software comes with a 54-page illustrated user's manual and a 5-1/4-inch disk for $49.95 plus $2.00 postage. (If you don't have an Apple, you can still benefit from "Annie's" analysis. For a small charge, an analysis of your antenna or array will be performed and the results mailed back to you.)

For more information, contact Sonnet Software, Dept. HR, 4397 Luna Course, Liverpool, New York 13088.

Circle #310 on Reader Service Card.

CAT technology

A new generation of CATs (computer aided transceivers) — the FT-757GX line — is available from Yaesu Electronics.



Controlled by three 8-bit microprocessors, the FT-757GX is a full QSK synthesized transceiver offering general coverage on receive and ham band transmit capability, with expanded coverage available for MARS operators. The transmitter section is specified for up to 30 minutes of continuous operation at a nominal output of 100 watts. For maximum operating flexibility, the FT-757GX performance package includes dual VFOs, eight memories, all-mode squelch, and a variety of scanning features. A 600 Hz CW filter, electronic keyer module, AF speech processor, and FM capability are all included in the purchase price.

Among the high-performance options available for the FT-757GX line are the FC-757AT automatic antenna tuner with band/antenna memory, the FP-757GX compact switching regulator power supply, the FP-757HD heavy-duty power supply (for continuous duty applications), the FP-700 standard power supply, and the FTV-700 transverter.

For further information on the FT-757GX line or other Yaesu transceivers designed for computer interface, contact Yaesu Electronics Corporation, P.O. Box 49, Paramount, California 90723.

Circle #311 on Reader Service Card.

color-coded A/V cable boots

Cole-Flex has announced the availability of a new color-coded insulating connector boot designed to fit many standard connectors. These heat-shrinkable boots are designed for use on audio and video cables and connectors to provide quick visual identification and interconnection. The boots also provide increased strain relief for cable assemblies and extra protection of connectors from rough handling.

Constructed of irradiated polyolefin, the boots have a 2:1 shrink ratio and can be installed with any standard industrial heat gun. They are available in a standard 1-3/4-inch length and can cover a standard 3/4-inch audio connector such as an XLR type. Colors available include black, white, red, yellow, blue, and clear. Hot stamping with special lettering or logos is also available.

For details, contact Cole-Flex Corporation, 91 Cabot Street, West Babylon, New York 11704.

Circle #312 on Reader Service Card.

Dressier 2-meter amplifier

The typical 2-meter transceivers available today often have noise figures of 6 to 8 dB. Adding a low-noise preamplifier ahead of the receiver will usually make a dramatic improvement in sensitivity. However, we must also consider the typical installation with a 1 to 2 dB transmission line loss from the antenna system, which further attenuates signals even before they reach the preamplifier. In particular, OSCAR 10 users have been experiencing such problems hearing downlink signals since this satellite may be up to 23,000 miles (38,000 km) distant instead of 900 miles (1500 km) as previous satellites were.

Several low-noise preamplifiers have recently become available, but they are primarily meant to be installed directly in the receiver at the shack and hence cannot overcome feedline losses. Dressler of West Germany has solved this problem by designing a preamplifier that can not only be easily mounted at the antenna but also remotely bypassed so that you can transmit around it with up to 1000 watts PEP.

This preamplifier is built on a high quality glass epoxy board and housed in a sealed metal enclosure. It is mounted in a waterproof ABS plastic outer housing complete with an adjustable clamp that will accommodate a mast up to 2 inches (51 mm) in diameter. Input/output connectors are type "N." The preamplifier features a low-noise dual gate GaAs (gallium arsenide) FET followed by a low-noise J-FET, thus yielding excellent noise figure as well as dynamic range. The input tank is a silver plated inductor. The overall bandwidth of the amplifier is filtered to 4 MHz so as to not respond to out of band signals, etc. Low insertion loss, high isolation coaxial relays perform the bypassing function and are mounted right on the preamplifier circuit board in order to take advantage of the low noise figure of the devices and protect them from burnout during transmit. Protection diodes are also

NEW products

located on both the input and output of the preamplifier to further protect the circuitry.

Its performance is outstanding. The measured noise figure on an HP 8970A noise figure generator was below 0.75 dB and gain is typically 18 dB. Typical receive operating current is 180 mA at a nominal 12 volts. No power is required during transmit. To energize the preamplifier, connect 12 to 15 volts to the feedthrough capacitor on the housing. For remote powering, the supply voltage may be applied directly to the antenna feedline with a bias inserter or through the Dressler model VV-INTERFACE remote power unit.

A 70 cm (432 MHz) unit will be available shortly. The 2-meter units are available in the United States from International Media Service, Box 26, Tewksbury, Massachusetts 01876. Price on the model EVV-2000 GaAs preamplifier is $109.95 plus $5 shipping; the model VV-INTERFACE is $29.95 plus $2.50 shipping.

Circle #313 on Reader Service Card.

cordless phone protection

The Kleen Line Protection System for cordless phone base stations is designed to suppress damaging telephone and power line spikes caused by lightning, spherics or phone office switch gear. The security system uses modern semiconductor, gas discharge tube and metal oxide varistor suppression techniques. Model PDS-11/SUP, priced at $81.95, has suppression on red and green phone lines (pins 3 and 4), with yellow and black lines brought straight through. A 6500 ampere suppressor protects the AC power line. Standard modular 4-pin telephone connectors provide simple, trouble-free hook-up.

For details, contact Electronic Specialists, Inc., 171 South Main Street, P.O. Box 389, Natick, Massachusetts 01760.

Circle #314 on Reader Service Card.

new satellite receiver

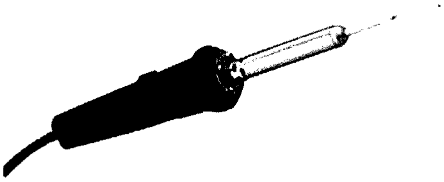
A new satellite receiver — the System 70 — features detent tuning, polarity control, a signal strength meter, a built-in modulator, scan tuning and both wide and narrow audio filters. It is available in two versions, the standard model 70X or the stereo version, model 70S, which decodes both matrix and discrete stereo sound and features simplified stereo tuning. Both models carry a full one year warranty.

For details, contact Lowrance Electronics, Inc., 12000 E. Skelly Drive, Tulsa, Oklahoma 74128.

Circle #315 on Reader Service Card.

dual heat soldering iron

The new Archer® Switchable Dual Heat Soldering Iron (No. 64-2055) from Radio Shack, a division of Tandy Corporation, completes small or medium-size jobs that would normally require two separate irons.



A convenient handle-mounted wattage switch allows the user to select 15 or 30 watts, depending on the size of the job, with no need to change the iron. The sculptured handle assures comfortable soldering. The U.L. listed dual-heat iron, 8-1/4-inches long, is available at participating Radio Shack Stores and sells for $6.95. Replacement tips (No. 64-2065), rated up to 30 watts, are priced at 89 cents each.

For information, contact Tandy Corporation/Radio Shack, 1800 One Tandy Center, Fort Worth, Texas 76102.

Circle #316 on Reader Service Card.

battery/button cell tester

While there are plenty of battery testers on the market, not all will test button cells — and some won't even accommodate the popular "N" size battery. The new CEC-1 Battery Tester and Button Cell Checker from Century Electronics tests all standard sizes plus all button types. Special compartments accommodate each battery size, and all batteries are automatically connected across a load resistor, for accurate readings on the unit's colorful, easy-to-read scale. The CEC-1 is priced at $7.95.

For more information, contact Century Electronics Corporation, 3511 North Cicero Avenue, Chicago, Illinois 60641.

Circle #317 on Reader Service Card.

buyers' guide

The Buyers' Guide To Radio And Electronic Parts is written so the hobbyist can easily locate a wide variety of electronic and mechanical parts for construction projects. It lists radio, antenna, computer, microwave, electronic, and mechanical parts sold by over 70 companies.

The Guide has two main sections: a directory of parts and a supplier information section. The directory lists the parts alphabetically by generic name with a part number and a description. A number is included on each line that tells the user the variety of parts that are stocked by the supplier for that listing, and is an indication of how complete a supplier's offering is for that part.

NEW products

Once the desired part is found in the directory and a supplier chosen, the user refers to the supplier information section, where company information is given. Included is the supplier's address and phone number, cost to obtain a catalog, minimum order information, and whether their parts are new or surplus. (To be included in the Guide, a supplier must have a mail order operation and be willing to sell in small quantities to individuals.)

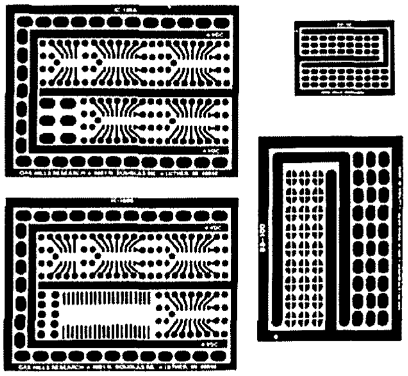
This book is available for $6.95 postpaid from Ham Radio's Bookstore, Greenville, New Hampshire 03048.

Circle #318 on Reader Service Card.

breadboards

The hard work of converting a raw piece of circuit board into a breadboard has been eliminated by Oak Hills Research with its new line of G-10 epoxy breadboards.

If soldering temperatures are kept low, these boards can be used repeatedly. (A 30-or 40-watt maximum rating is recommended for the pencil iron used; in many instances a 25-watt iron will be sufficient.) They are also excellent for finished, permanent projects. Double-sided versions of these boards are available on special order for those who require a ground plane surface opposite the etched foils.



Two general-purpose breadboards are available. They are the BB-50 for small projects (2-1/4 × 1-3/4 inches) and BB-100 (3-1/4 × 4-3/4 inches). While both units have the same pattern, the BB-100 includes isolated pads on one side of the board divided into four equal segments to accommodate transistors with as many as four leads.

Two IC breadboards are also available. They are the IC-100A (4 × 5-1/4 inches) and IC-100B (4 × 5-1/4 inches). The 100A contains sites for four 16-pin DIP ICs, plus one 8-pin DIP. Numerous isolated pads are included for mounting additional components. The IC-100B will accommodate three 16-pin DIP ICs, one 8-pin DIP

and an LSI chip with as many as 40 pins. Each IC site is marked for drilling, should the user desire to install IC sockets.

All feature ground and plus voltage buses routed through to provide short connections to these conductor lines. The outer borders of the boards are copper, and they are common to the ground bus of the board. This aids grounding when mounting the breadboards to a metal mainframe. All conductors are tin plated to retard corrosion.

Prices are: BB-50, $2; BB-100, $3; IC-100A and B, $4. (Include $1 for handling and postage for orders under $10; postage on orders greater than $10 is paid by Oak Hills Research. Michigan residents include sales tax.)

For information or orders, contact W1FB, Oak Hills Research, 4061 N. Douglas Road, Luther, Michigan 49656.

Circle #319 on Reader Service Card.

new Larsen® catalog

The redesigned, simplified-format Larsen Electronics catalog offers more thorough explanations of products, electrical types, and frequencies, thereby making ordering easier. Now available, the catalog includes all of the latest developments in both antennas and accessories at prices unchanged since August, 1982.

For more information, contact Larsen Electronics, P.O. Box 1799, Vancouver, Washington 98668.

Circle #320 on Reader Service Card.

beacon guide

The new edition of *The Beacon Guide* — updating the first edition, published in 1974 — is now available by mail or through selected retailers. Compiled by H. John Clements, WA6RXN, and edited by Ken Stryker, the volume is distributed exclusively by the Century Print Shop, 6059 Essex Street, Riverside, California 92504. The price of the 100-page book is $7; club and other quantity discounts are available.

For information, contact the Century Print Shop, 6059 Essex Street, Riverside, California 92504.

Circle #321 on Reader Service Card.

CMOS keyer kit

BEL-TEK has introduced an inexpensive CMOS keyer kit incorporating state-of-the-art circuitry. Power consumption is kept to a minimum by the use of low power CMOS digital integrated circuits.

The keyer uses a triggered clock to completely eliminate the possibility of having the first dot or dash any longer than the characters that follow. The instant you push the key, the character starts, eliminating the delay often encountered in keyers with free-running clocks. The digital circuitry of the CMOS keyer provides an exact 3:1 weight ratio for perfect CW. The keyer also provides jam-proof spacing which eliminates

the chance of placing dots and dashes too close together. After a dot or dash is completed, a space is automatically inserted before the next character begins, even if the key was pushed before the completion of the space. If both the dot and dash keys are sent simultaneously, the dash will dominate until it is released.

The keyer can be powered by any voltage between 5 and 12 volts DC. The circuit is protected against accidental polarity reversal of the supply voltage. The keyer can be operated at any speed between 5 and 50 WPM by adjusting the speed control to your desired operating speed. The built-in 800 Hz sidetone has an adjustable volume control.

A rugged high voltage keying transistor is used to eliminate annoying problems which plague mechanical relays. The transistor will reliably switch loads up to 250 mA. The keyer is compatible with grid block, cathode keyed, and solid-state transmitters.

The CMOS keyer kit comes with a printed circuit board and all components necessary for assembling. The price is $9.95 plus $1.50 shipping.

For more information, contact BEL-TEK, P.O. Box 125H, Beloit, Wisconsin 53511.

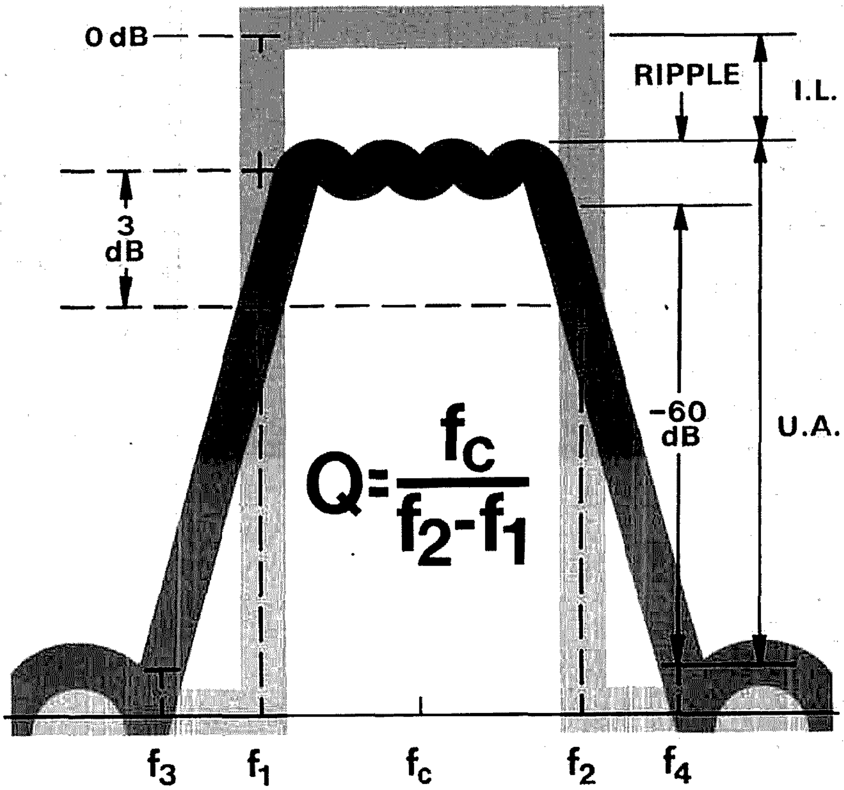
Circle #322 on Reader Service Card.

APRIL 1984 / $2.50

NEW: SHORT CIRCUIT HOTLINE — see page 6

ham radio magazine

- a portable SW receiver
- homebrew spectroscope
- graphic filter design
- master the Morse keyboard
- PL tone generator
- high-voltage power supply
- understanding the interface loop
- branch-line hybrid



RESONANT CIRCUITS EXPLAINED

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contents

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Welcome to the Hotline

Have you ever read a construction article in a magazine, then become all excited about putting the equipment together — spending weeks searching through junk boxes, flea markets, and your local radio outlet for all the parts — only to find that when you sat down and put it all together it didn't work? Then, when you dug into the circuit and made sure it was wired correctly — per the article — you found an error in the schematic? Needless to say, your feelings at times like this are nothing *short of unprintable.*

All publishers, *ham radio* included, have a difficult problem in this area. We work extremely hard to catch any errors as our material is prepared for publication. Unfortunately, no matter how much effort is put into error control, problems sometimes occur. To further compound the problem, it takes at least two or three months to publish a correction; by the time the correction finally appears in print, a number of people have been inconvenienced.

After giving some serious thought to this whole matter we realized we could improve this situation, at least for the readers of *ham radio.* So as of April 1st we'll be operating a special "Ham Radio Short-Circuit Hotline" that you can call from 9 AM to Noon and 1 to 3 PM, Eastern Time, Monday through Friday. Here you can find out whether there are any last-minute changes or problems you should be aware of in your particular project *before* you start work on it. The Hotline number is **603-878-1441**. (No collect calls, please.)

As an additional service, we'll keep your name in a file of readers we know are working on a particular project. If any problems should arise after your call, we'll drop you a line telling you what changes are involved.

In using this service, please remember that the Hotline isn't a technical advisory service. You'll be talking with a clerk, not a technically trained Amateur. *Please don't call the Hotline for help with construction; if you have a problem, write to the author of the article in question.*

And if you find an error in an article, tell us in a brief letter. We'll check it out, and if we find that there is indeed an error, we'll add it to our Hotline file and prepare a "Short Circuit" for publication in the next available issue.

It's important to remember that this service will cover only problems that would render the project either unusable or unsafe. We regret that we cannot act as a clearing house for design ideas or offer advice on how to squeeze the last dB of performance out of the equipment concerned. If you have comments of that nature, we'd welcome them; just jot them down and address them to Rich Rosen, Editor-in-Chief, in a letter.

Hopefully, this will be a good step toward helping to solve this familiar problem and will make *ham radio* even more valuable to all *of you* builders.

Skip Tenney, W1NLB
Publisher

presstop de W9JUV

APPOINTMENTS FOR AMATEUR EXAMS ARE FILLED UNTIL AUGUST in many FCC Field Offices, as a result of the drastic cutback in FCC-administered exams effective January 1. Word about the new four-exam-per-year schedule got around very quickly following the first session in February, and both potential new Amateurs and present licensees wishing to upgrade have already filled all available slots in many offices' May session. The effect is also being seen in the number of new licenses being issued, which dropped drastically in late January after applications from December testing were processed by Gettysburg.

FIVE VOLUNTEER EXAMINER COORDINATORS HAVE NOW BEEN AUTHORIZED by the FCC, with VEC agreements with groups in Alaska, Puerto Rico, and the second, eighth, and ninth call areas all signed February 23. Formal acceptance of all five had been delayed by internal FCC procedural problems, not unexpected when such a far-reaching new program goes into effect for the first time. However, even though the new VECs have been formally accepted it will still take some time for them to get their programs on line. Serious efforts toward setting up VECs in the fifth and sixth call areas are also reported under way, but the remaining districts seem to be a long way from an adequate testing schedule.

ARRL Progress In The VEC Program Seems To Be At A Standstill, with the directors still agreeing that they won't proceed any further until the FCC agrees to permit the collection of a fee. In addition, it appears likely that the ARRL doesn't want to be involved in the exam program at all unless it is on a national basis—despite the FCC's report and order that VECs be regional though permitting one VEC to serve in more than one district.

FCC's NPRM To Permit Amateur Exam Fee Collection Is Due momentarily, and may well be out before you read this. Key questions the FCC would like addressed in responses to its Notice of Proposed Rule Making include what portion (if any) of the fee should go to the examiner, and who should collect the fee. The NPRM may have a very short Comment time, so those wishing to file comments should be prepared to respond rather quickly.

RE-EXAMINATION FOR CODE PROFICIENCY EVERY TWO YEARS is being sought in a Petition for Rule Making submitted by Wayne Green, W2NSD. In addition, he also proposes requiring five wpm upgrades at each two-year interval, to a final level of 35 wpm. Failure to improve CW speed would lead to loss of license. At presstime no RM number had been assigned.

Novice Phone Privileges On 220 MHz Have Been Proposed in another Petition for Rule Making filed by WA2MCT/5. He'd permit simplex voice and RTTY operation by entry-level licensees from 223.4 to 223.95 MHz. Such a move would, he feels, help protect the 220-MHz band from developing pressures by commercial services. It has no RM number yet, either.

COMMUNICATIONS FOR THE U.S. 1984 OLYMPIC TORCH RUN will be provided by Amateur Radio. The run, which begins May 8 on the East Coast and will cover 13,000 miles and all the 48 contiguous states before ending July 28 in Los Angeles, will be coordinated by Amateurs operating from stations in the fleet of support vans. AT&T is supporting the event, and the Amateur operators will be AT&T or subsidiary employees donating their vacation time.

The FCC Is NOT Directly Involved With Olympic Communications, Amateur or otherwise, at this summer's Olympics. Amateurs with questions regarding Olympic communications and Amateur Radio should contact W6EJJ or K6ZT. It presently appears that, because of security considerations, there may be no Amateur Radio operation from the Olympic village as there has been in previous Olympics. The FCC does plan a strong, well coordinated effort against jammers or other radio abusers on Amateur bands and other frequencies during the Olympics.

THE FORMAL AGREEMENT BETWEEN ARRL AND REACT to facilitate cooperation between the two groups during emergency situations may now be a dead issue. The effort, which had begun last May by the ARRL board, was tabled at the October board meeting in Houston, reportedly out of concern for the results of CB delicensing. Despite a letter from REACT Executive Director Gerald Reese noting that the agreement would be with the REACT organization and not individual CB operators, and that it would simply recognize officially a mutually beneficial arrangement that already exists in many parts of the country, it now appears the board does not plan to proceed with an agreement in the foreseeable future.

QRM TO AMATEURS ON 70 CM HAS RESULTED IN THE SHUTDOWN of coastline navigation devices in California and may bring similar action along the Gulf Coast. Navigation Services, Inc., turned off eight 433 and 437 MHz transmitters in Southern California after the FCC confirmed their operation was causing harmful interference to area fast-scan ATV users.

The Gulf Coast Situation Is Similar, though in that case an attempt to arrange for "peaceful coexistence" is being made with the navigation system operator, Sercel. Amateurs do share the 70-cm band with other users, but on a strict non-interference basis.

"THE AMSAT-STONER CHALLENGE CUP," an OSCAR 10 operating competition, celebrates the 25th anniversary of W6TNS's April, 1959 proposal in CQ magazine that there could be an Amateur communications satellite. Competitors are to work (or for SWLs, hear) as many grid squares as possible from April 15 to July 14, 1984. First place winner will receive a silver cup—other top winners, plaques. Complete rules appear in Amateur Satellite Report and Orbit magazine. A contest "package" that includes copies of the historic articles that led to the contest is available from AMSAT for $2.

resonant circuits

Understanding the relationship between component and circuit Q and source and load impedances

How well do you understand the basic resonant circuit? In this excerpt from his recent book, *RF Circuit Design*, Chris Bowick explains the fundamentals of RF circuitry.

The resonant circuit is certainly nothing new in Amateur Radio. Used in practically every transmitter, receiver, or piece of test equipment in existence, it selectively passes a frequency or group of frequencies from a source to a load while attenuating all other frequencies outside of this passband.

The ideal resonant circuit passband, as pictured in **fig. 1**, would be rectangularly shaped, with infinite attenuation above and below the frequency band of interest. No attenuation would be introduced at the signal frequency. But the realization of this filter is, of course, impossible because of the physical characteristics of the components which make up the filter; because there is no such thing as a perfect component, there can be no perfect filter. However, understanding the mechanics of resonant circuits enables us to tailor an imperfect circuit to suit our needs.

Fig. 2, a more realistic representation, shows what a practical filter response might resemble. Applicable definitions are presented below:

The **bandwidth** of any resonant circuit is commonly defined as the difference between the upper and lower frequency $(f_2 - f_1)$ of the circuit at which its amplitude response is 3 dB below the passband response. It is often called the half-power bandwidth.

The ratio of the center frequency of the resonant circuit to its bandwidth is defined as the *circuit-Q.*

$$Q = \frac{f_c}{f_2 - f_1} \tag{1}$$

This Q should not be confused with *component Q* which is defined as the ratio of a component's reactance (X) to its series resistance (R_S). While component Q does have an effect on circuit Q, the reverse is not true; circuit Q is a measure of the selectivity of a resonant circuit. The higher its Q, the narrower its bandwidth, and thus, the higher its selectivity.

The **shape factor** of a resonant circuit is typically defined as the ratio of the 60 dB bandwidth to the 6 dB bandwidth. For example, with a 60 dB bandwidth of 3 MHz and a 6 dB bandwidth of 1.5 MHz the shape factor is:

$$SF = \frac{3\ MHz}{1.5\ MHz} = 2$$

Shape factor is a measure of the steepness of the skirts; the smaller the number, the steeper the response. Notice that our perfect filter of **fig. 1** has a shape factor of 1, which is the ultimate. The passband for a filter with a shape factor *smaller* than 1 would have to look similar to the one shown in **fig. 3**. Obviously, this is a physical impossibility.

Ultimate Attenuation, as the name implies, is the final *minimum* attenuation that the resonant circuit presents outside of the specified passband. A perfect resonant circuit would provide infinite attenuation outside of its passband. However, with real components, infinite attenuation is impossible to obtain. (Keep in mind that if the circuit presents response peaks outside the passband, as shown in **fig. 2**, then this, of

*Adapted with permission from *RF Circuit Design* by Chris Bowick, Howard W. Sams & Co., Indianapolis, Indiana 46206. (Available from Ham Radio's Bookstore, Greenville, NH 03048, $24.95 postpaid.)

By **Chris Bowick, WD4C**, 200 Abri Place, Lilburn, Georgia 30247

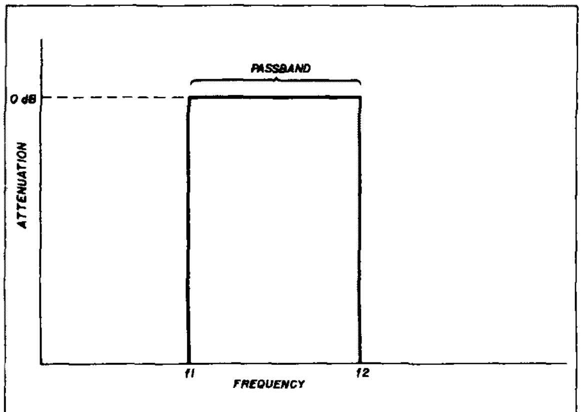

fig. 1. Ideal filter response.

course, will detract from the ultimate attenuation specification of that resonant circuit.)

Insertion Loss occurs whenever a component or group of components is inserted between a generator and its load, and some of the signal from the generator is absorbed in those components as a resistive loss. The resulting attenuation, called insertion loss, a very important characteristic of resonant circuits, is usual*ly expressed in decibels* (dB).

Ripple, a measure of the flatness of the passband of a resonant circuit, is also expressed in dB. It is the difference between the maximum and minimum attenuation *in the passband.*

resonance (lossless components)

What is resonance? What causes it to occur,and how can it be used to best advantage? Using voltage division, whenever a shunt element of impedance Z_p is placed across the output of a generator with internal resistance R_s, **fig. 4**, the maximum output voltage available from this circuit is:

$$V_{out} = \frac{Z_p}{R_s + Z_p} (V_{in}) \qquad \textbf{(2)}$$

V_{out} is always less than V_{in}. If Z_p varies with frequency, such as would occur with capacitive or inductive reactance, then V_{out} is also frequency dependent and the ratio of V_{out} to V_{in}, which is the gain (or in this case, loss) of the circuit, is also frequency dependent. Let's take, for example, a 25 pF capacitor as the shunt element (**fig. 5A**) and plot the function V_{out}/V_{in} in dB versus frequency, where we have:

$$\frac{V_{out}}{V_{in}} = 20 \log_{10} \frac{X_c}{R_s + X_c} \qquad \textbf{(3)}$$

where $\frac{V_{out}}{V_{in}}$ = loss in dB

R_s = source resistance

X_c = capacitive reactance $= \frac{1}{j\omega C}$

The plot of this equation is shown in **fig. 5B.** Note that loss increases as the frequency increases; thus we have formed a simple *lowpass filter.* Notice also that the attenuation *slope* eventually settles down to the rate of 6 dB for every octave (doubling) increase in frequency (also called 6 dB "roll-off"). This is due to the single reactive element in the circuit. This attenuation slope increases an additional 6 dB for each *significant* reactive element that we insert into the circuit.

If we now delete the capacitor from the circuit and insert a 0.05 μH inductor in its place, we obtain the circuit of **fig. 6A** and the plot of **fig. 6B**, where

$$\frac{V_{out}}{V_{in}} = 20 \log_{10} \frac{X_L}{R_s + X_L} \qquad \textbf{(4)}$$

where $\frac{V_{out}}{V_{in}}$ = loss in dB

R_s = source resistance

X_L = coil reactance $= j\omega L$

Here we have formed a simple *highpass filter* with an attenuation slope of 6 dB per octave.

Simple calculations using the basic voltage division formula (**eq. 2**) enabled us to plot the frequency response of two separate and opposite reactive components. But what happens if we place both the inductor and capacitor across the generator simultan-

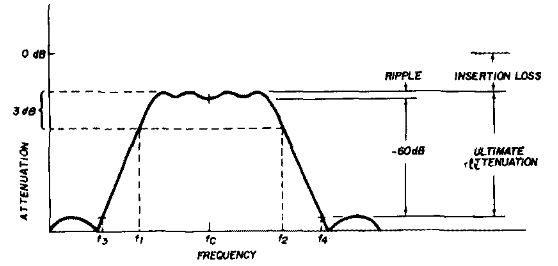

fig. 2. A practical filter response.

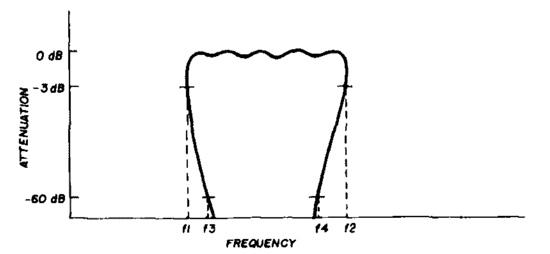

fig. 3. An impossible filter shape factor.

eously? This case is no more difficult to analyze than the previous two circuits. In fact, at any frequency, we can simply apply the basic voltage division rule as before. The only difference here is that we now have two reactive parallel components to deal with instead of one. The circuit is shown in **fig. 7A** and its response is plotted in **fig. 7B**. Notice that as we approach the resonant frequency of the tuned circuit, the slope of the resonance curve increases. This occurs because two *significant* reactances are present and each one is changing at the rate of 6 dB per octave, with slopes in opposite directions. As we move away from resonance in either direction, however, the curve again approaches a 6 dB/octave slope because again only one reactance becomes significant. The other reactance presents a very high impedance to the circuit at these frequencies and the circuit behaves as if that reactance were no longer there.

loaded-Q

The Q of a resonant circuit was defined earlier to be equal to the ratio of the circuit's center frequency to its 3 dB bandwidth (**eq. 1**). This "circuit Q," as it was called, is often given the label "loaded-Q" because it describes the passband characteristics of the resonant circuit under actual in-circuit or *loaded* conditions. The loaded-Q of a resonant circuit is dependent upon three main factors (see **fig. 8**): (1) the source resistance (R_S), (2) the load resistance (R_L), and (3) the component Q.

effect of R_S and R_L on loaded Q

The role that source and load impedances play in determining the loaded Q of a resonant circuit is probably best illustrated through an example. In **fig. 7** we plotted a resonance curve for a circuit consisting of a 50-ohm source, a 0.05 μH lossless inductor, and a 25 pF lossless capacitor. The loaded Q of this circuit, as defined by **eq. 1** and determined from the graph, is approximately 1.1. Obviously, this is not a very narrow band or high-Q design. However, after replacing the 50 ohm source with a 1000 ohm source and again plotting the results, the response in **fig. 9** is applicable. (The resonance curve for the circuit with the 50 ohm source is shown in dashed lines for comparison. Notice that the Q, or selectivity of the resonant circuit, has increased dramatically to approximately 22. Thus, by raising the source impedance, we have increased the Q of our resonant circuit.

However, neither of these plots addresses the effect of the load impedance on the resonance curve. If an external load were attached to the resonant circuit, as shown in **fig. 10A**, the effect would be to broaden or "de-Q" the response curve, depending

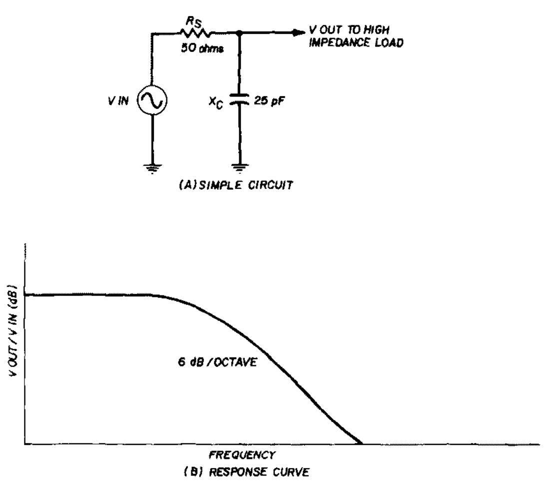

fig. 5. Frequency response of a simple RC lowpass filter: *(A)* shows simple circuit; *(B)*, response curve.

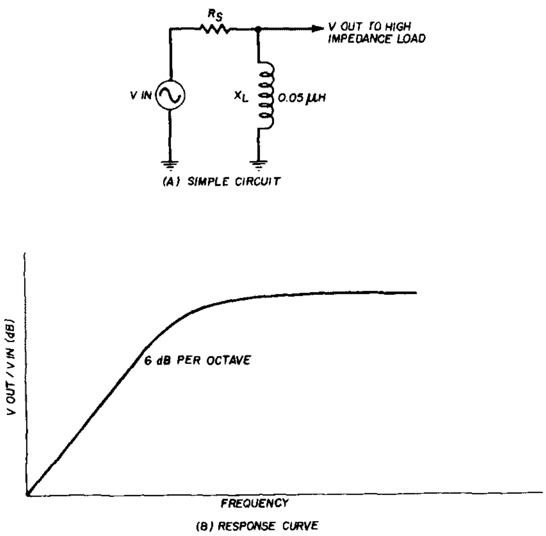

fig. 6. Simple highpass filter: *(A)* shows simple circuit; *(B)*, response curve.

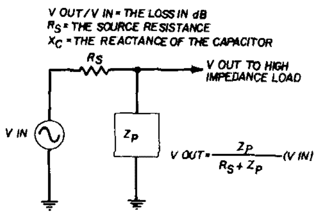

fig. 4. Voltage division rule.

upon the value of the load resistance. The equivalent circuit is shown in **fig. 10B**. The resonant circuit sees an equivalent resistance of R_s in parallel with R_L as its true load. Because this total external resistance is, by definition, smaller in value than either R_s or R_L, the loaded Q must decrease. That is, assuming lossless components,

$$Q = \frac{R_p}{X_p} \tag{5}$$

where R_p = equivalent parallel resistance of R_s and R_L
X_p = either the inductive or capacitive reactance (they are equal at resonance)

Eq. 5 illustrates that a decrease in R_p also decreases the Q of the resonant circuit and an increase in R_p increases the circuit Q. It also illustrates another very important point: the *same* effect can be obtained by keeping R_p constant and varying X_p. Consequently, for a given source and load impedance, the optimum Q of a resonant circuit is obtained when the inductor is a small value and the capacitor is a large value; in either case, X_p is decreased. This effect is shown in **figs. 11** and **12**.

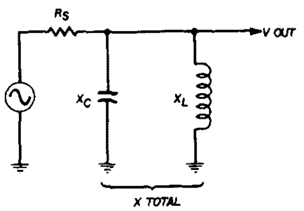

fig. 7A. Resonant circuit with two reactive components.

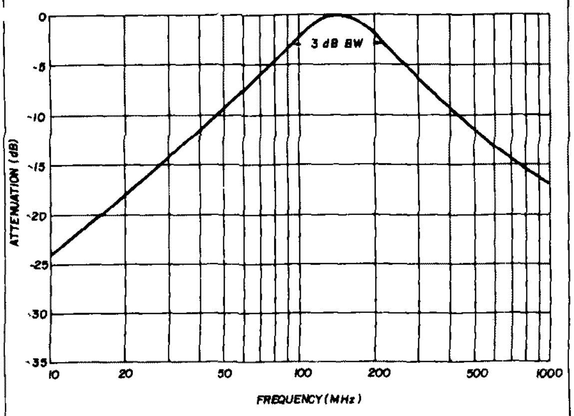

fig. 7B. Frequency response of an LC resonant circuit.

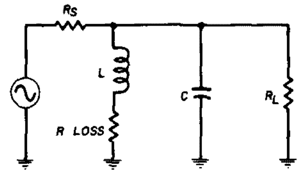

fig. 8. Circuit for loaded-Q calculations.

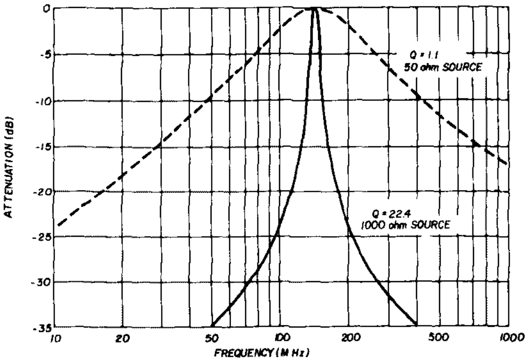

fig. 9. The effect of R_s and R_L on loaded Q.

Two possible approaches to designing a resonant circuit with a particular Q are available: selecting an optimum value of source and load impedance or selecting L and C component values which optimize Q. Often there is no real choice because in many instances the source and load are defined and cannot be varied. When this occurs, X_p is automatically defined for a given Q and we usually end up with component values that are impractical at best. Methods of eliminating this problem will be shown.

example 1

Design a resonant circuit to operate between a source resistance of 150 ohms and a load resistance of 1000 ohms. The loaded Q must be equal to 20 at the resonant frequency of 50 MHz. Assume lossless components and no impedance matching.

Solution: The effective parallel resistance across the resonant circuit is 150 ohms in parallel with 1000 ohms, or

$$R_p = 130 \text{ ohms}$$

Using **eq. 5**

$$X_p = \frac{R_p}{Q} = \frac{130}{20} = 6.5 \text{ ohms}$$

and $$X_p = \omega L = \frac{1}{\omega C}$$

therefore $$L = 20.7 \text{ nH}$$

$$C = 489.7 \text{ pF}$$

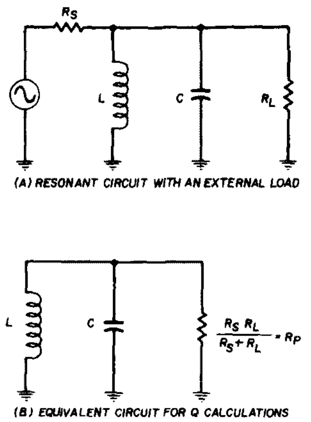

fig. 10. The equivalent parallel impedance across a resonant circuit: *(A)* shows resonant circuit with an external load; *(B)*, equivalent circuit for Q calculations.

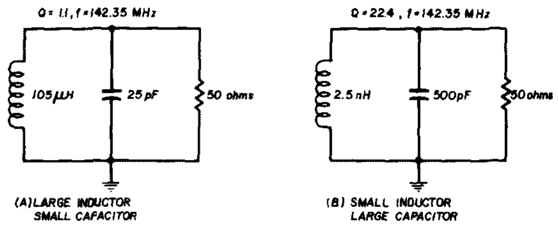

fig. 11. Effect of Q versus X at 142.35 MHz: *(A)* shows large inductor, small capacitor; *(B)*, small inductor, large capacitor.

the effect of component Q on loaded Q

Up to this point we have assumed that the components used in the resonant circuits are lossless and do not degrade the loaded Q. In reality, however, such is not the case; individual component Q's must be taken into account. In a lossless parallel resonant circuit, the impedance seen across its terminals at resonance is infinite. In a practical circuit, however, due to component losses, some finite equivalent parallel resistance exists, as shown in **fig. 13**. This resistance (R_p) and its associated shunt reactance (X_p) can be found from the following transformation equations:

$$R_p = (Q^2 + 1) R_s \qquad \textbf{(6)}$$

where R_p = equivalent parallel resistance

$Q = Q_s = Q_p =$ *component Q*

R_s = component series resistance

and $$X_p = \frac{R_p}{Q_p} \qquad \textbf{(7)}$$

If the Q of the component is greater than 10, then

$$R_p \approx Q^2 R_s \qquad \textbf{(8)}$$

and $$X_p \approx X_s \qquad \textbf{(9)}$$

These transformations are valid at only one frequency because they involve the component reactance which is frequency dependent. The following example illustrates this point.

example 2

Given the 50 nanohenry coil shown in **fig. 14A**, compute its Q at 100 MHz. Then transform the series circuit of **fig. 14A** into the equivalent parallel inductance and resistance circuit of **fig. 14B**.

Solution: The Q of this coil at 100 MHz is:

$$Q = \frac{X_s}{R_s}$$

$$= \frac{2\pi\,(100 \times 10^6)\,(50 \times 10^{-9})}{10} = 3.14$$

Since the Q is less than 10, **eq. 6** is used to find R_p.

$$R_p = (Q^2 + 1)\,R_s$$

$$= [(3.14)^2 + 1]\,10 = 108.7 \text{ ohms}$$

Next, we find X_p using **eq. 7**.

$$X_p = \frac{R_p}{Q} = \frac{108.7}{3.14} = 34.62$$

and, the parallel inductance becomes

$$L_p = \frac{X_p}{\omega}$$

$$= \frac{34.62}{2\pi(100 \times 10^6)} = 55.1 \text{ nH}$$

These values are shown in **fig. 14B**.

Example 2 clearly illustrates the undesirable effects of using low Q components in highly selective resonant circuit designs; doing this is equivalent to placing a low value shunt resistor directly across the circuit. As we saw in an earlier section, any low-value resistance shunting a resonant circuit drastically reduces its loaded Q and increases its bandwidth.

In most cases, we only need to know the Q of the *inductor* in loaded Q calculations. The Q of most capacitors is quite high over their useful frequency range, and the equivalent shunt resistance they pre-

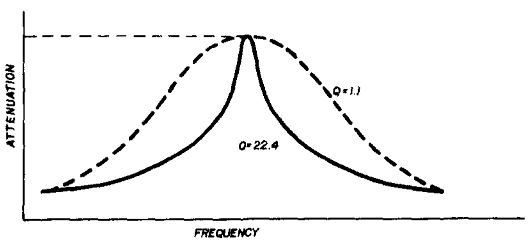

fig. 12. Plot of loaded-Q curves for circuits in *fig. 11*.

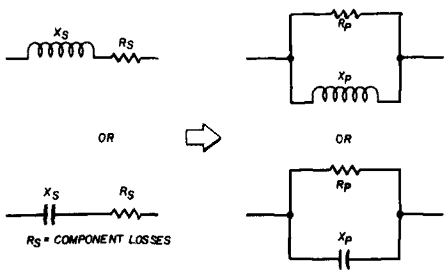

fig. 13. A series-to-parallel transformation.

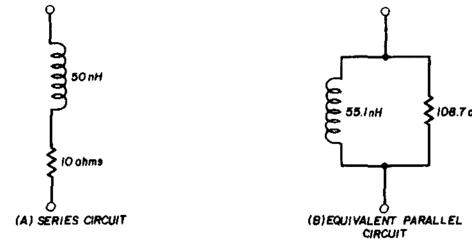

fig. 14. Example of a series-to-parallel transformation: ***(A)*** **shows series circuit;** ***(B)*****, shows equivalent parallel circuit.**

sent to the circuit is also quite high, and can usually be neglected. Care must be taken, however, to ensure that this is indeed the case.

insertion loss and component Q

If inductors and capacitors were perfect, rather than lossy, then insertion loss for L-C resonant circuits and filters would not exist. This is, of course, not the case and as it turns out, insertion loss is a very critical parameter in the specification of any resonant circuit.

Fig. 15 illustrates the effect of inserting a resonant circuit between a source and its load of equal value. In **fig. 15A**, the source is connected directly to the load. Using voltage division, we find that

$$V_l = 0.5\ V_{in}$$

Fig. 15B shows a resonant circuit placed between the source and the load. **Fig. 15C** is the equivalent circuit at resonance. Notice that the use of an inductor with a Q of 10 at the resonant frequency is equivalent to a shunt resistance of 4500 ohms. This resistance, combined with R_L, produces a 0.9 dB voltage loss at V_l in comparison with the equivalent point in **fig. 15A**.

An insertion loss of 0.9 dB doesn't sound like much, but it can add up very quickly if several resonant circuits are cascaded. **Example 3** may help to clarify things.

example 3

Design a simple parallel resonant circuit to provide a 3 dB bandwidth of 10 MHz at a center frequency of 100 MHz. The source and load impedances are each 1000 ohms. Assume the capacitor to be lossless. The Q of the available inductor is 85. What is the insertion loss off the network?

Solution: From **eq. 1** the required loaded Q of the resonant circuit is:

$$Q = \frac{f_c}{f_2 - f_1} = \frac{100\ MHz}{10\ MHz} = 10$$

To find the inductor and capacitor values to complete the design requires that we know the equivalent shunt resistance and reactance of the components at resonance.

From **eq. 7**

$$X_p = \frac{R_p}{Q_p}$$

where X_p = reactance of the inductor and capacitor at resonance
R_p = equivalent shunt resistance of the inductor
Q_p = Q of the inductor = 85

Thus $R_p = 85X_p$ **(10)**

The loaded Q of the resonant circuit is equal to

$$Q = 10 = \frac{R_{TOTAL}}{X_p}$$

where R_{TOTAL} = shunt resistance
$= R_p || R_s || R_L$
(three parallel values: R_p, R_s and R_L)

or $$Q = 10 = \frac{\frac{R_p(500)}{R_p + 500}}{X_p} \qquad \textbf{(11)}$$

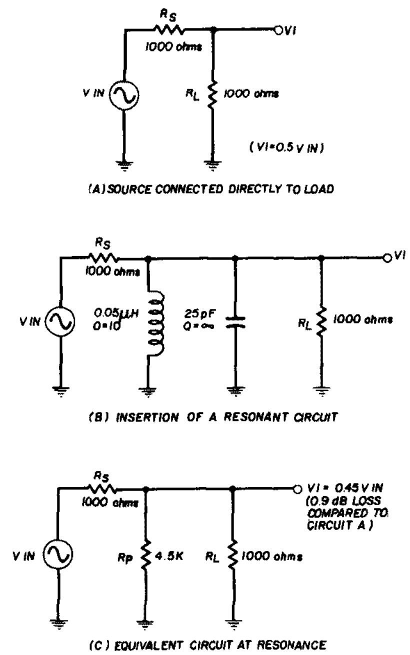

fig. 15. The effect of component Q on insertion loss: ***(A)*** **shows source connected directly to the load;** ***(B)*****, insertion of a resonant circuit;** ***(C)*****, equivalent circuit at resonance.**

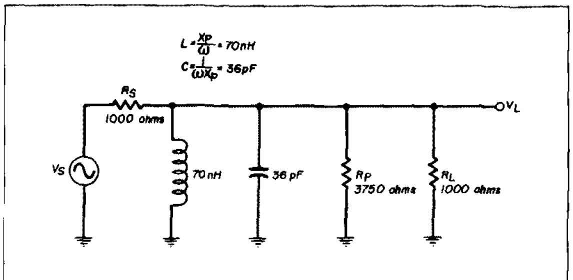

fig. 16. Resonant circuit design illustrates example 3.

We now have two equations and two unknowns (X_p, R_p). If **eq. 10** is substituted into **eq. 11**

$$X_p = 44.1 \text{ ohms}$$

Plugging this value back into **eq. 10** gives

$$R_p = 3.75 K$$

Thus, our component values are

$$L = \frac{X_p}{\omega} = 70 \, nH$$

$$C = \frac{1}{\omega X_p} = 36 \, pF$$

and the final circuit is shown in **fig. 16**.

The insertion loss calculation, at center frequency, is found by applying the voltage division rule as follows: R_p in parallel with R_L is equal to 789.5 ohms. The voltage at V_L is, therefore

$$V_L = \frac{789.5}{789.5 + 1000} (V_s)$$

The voltage at V_L without the resonant circuit in place is equal to $0.5 \, V_s$ due to the 1000 ohm load giving

$$\text{insertion loss} = 20 \log_{10} \frac{0.44 \, V_s}{0.5 \, V_s}$$

$$= 1.1 \, dB$$

impedance transformation

As we have seen, low values of source and load impedance tend to load a given resonant circuit down and thus decrease its Q and increase its bandwidth. This makes it very difficult to design a simple L-C high-Q resonant circuit for use between two very low values of source and load resistance. In fact, even if we were able to come up with a design on paper, it would most likely be impossible to build such a circuit because of the extremely small (or negative) inductor values that would be required.

One method of circumventing this problem is to make use of one of the impedance transforming circuits shown in **fig. 17**. These circuits increase the value of the source or load resistance presented to the resonant circuit. For example, an impedance transformer could present an impedance (R_s') of 500 ohms to the resonant circuit, when in reality we have an R_s of 50 ohms. By using transformers, the Q of the resonant tank and its selectivity can be increased. In many cases these methods can make a previously unworkable problem workable again, complete with realistic values for the coils and capacitors involved.

The design equations for each of the transformers are presented below and are useful for designs needing loaded Qs of greater than 10. For the tapped-C transformer (**fig. 17A**)

$$R_s' = R_s \left(1 + \frac{C1}{C2}\right)^2 \quad \textbf{(12)}$$

And the equivalent capacitance (C_T) that will resonate with the inductor is equal to $C1$ in series with $C2$ or

$$C_T = \frac{C1C2}{C1 + C2} \quad \textbf{(13)}$$

For the tapped-L network of **fig. 17B**

$$R_s' = R_s \left(\frac{n}{n_1}\right)^2 \quad \textbf{(14)}$$

Example 4 illustrates the use of these equations in a simple design.

example 4

Design a resonant circuit with a loaded Q of 20 at a center frequency of 100 MHz which will operate between a source resistance of 50 ohms and a load resistance of 2000 ohms. Use the tapped C approach and assume that the inductor Q is 100 at 100 MHz.

Solution: The tapped-C transformer is used to step the source resistance up to 2000 ohms to match the load resistance for optimum power transfer,

$$R_s' = 2000 \text{ ohms}$$

and from **eq. 12**, we have

$$\frac{C_1}{C_2} = \sqrt{\frac{R_s'}{R_s}} - 1$$

or $$C1 = 5.3 \, C2 \quad \textbf{(15)}$$

Proceeding as we did in **example 3**, we know that for the inductor

$$Q_p = \frac{R_p}{X_p} = 100$$

therefore

$$R_p = 100 \, X_p \quad \textbf{(16)}$$

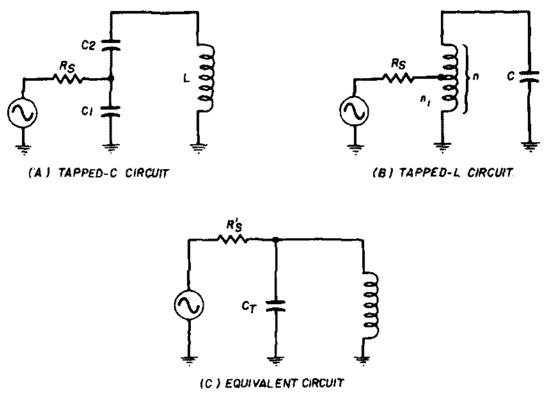

fig. 17. Two methods used to perform an impedance transformation: *(A)* shows tapped-C circuit; *(B)* tapped-L circuit. *(C)* shows equivalent circuit.

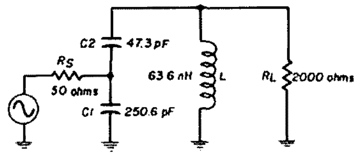

fig. 18. Final impedance transformation circuit.

We also know that the loaded Q of the resonant circuit is equal to

$$Q = \frac{R_{TOTAL}}{X_p}$$

where R_{TOTAL} = *total equivalent shunt resistance*

$$= R_s'||R_p||R_L = 1000||R_p$$

and where we have taken R_s' and R_L to each be 2000 ohms in parallel.

Hence the loaded Q is

$$Q = \frac{1000R_p}{(1000 + R_p)\,X_p} \tag{17}$$

Substituting **eq. 16** into **eq. 17** and solving for X_p yields

$$X_p = 40 \text{ ohms}$$

And substituting this result back into **eq. 16** gives

$$R_p = 4000 \text{ ohms}$$

and

$$L = \frac{X_p}{\omega} = 63.6 \text{ nH}$$

$$C_T = \frac{1}{X_p\omega} = 39.78 \text{ pF}$$

We now know what the total capacitance must be to resonate with inductor. We also know from **eq. 15** that $C1$ is 5.3 times larger than $C2$. Thus, if we substitute **eq. 15** into **eq. 13** and simultaneously solve the equations we get

$$C2 = 47.3 \text{ pF}$$
$$C1 = 250.6 \text{ pF}$$

The final circuit is shown in **fig. 18.**

As an exercise, you might want to rework **example 4** without the aid of an impedance transformer. You will find that the inductor value which results is much more difficult to obtain and control physically because it is so small.

conclusion

The parallel resonant circuit, as simple as it may seem, certainly offers quite a lot to think about. We've discovered why good quality, high-Q, components are desirable in any low-loss, narrowband, tuned circuit, and why lossy components (low-Q components) increase insertion loss and broaden the bandwidth of the circuit. We've also discovered the effects of source and load impedances on the loaded Q of the resonant circuit. More importantly, however, by understanding these effects we have mastered a method of controlling them through the use of impedance transformers.

ham radio

a handheld optical spectrum analyzer

Experiment with light build an inexpensive homebrew spectroscope

The history of Amateur and commercial radio, has been marked by the use of ever-increasing frequencies. Each era has seen applications found for the previously useless short wavelengths. With the introduction of lasers, communication in the optical frequency band is becoming attractive. Long distance fiber optic links use semiconductor laser diodes to achieve bandwidths up to several Gigahertz. Amateurs have communicated using gas and semiconductor lasers over moderate paths, and as the price of lasers continues to decline, more and more applications will be found — particularly by Amateurs in their role as experimenters and innovators in communications technology. But lasers are not the only usable source of light. Incandescent lamps, fluorescent lamps and especially light emitting diodes can and will continue to be used to transmit data.

basic spectrum analyzer

Most people do not consider light to be in the same class of phenomena as radio signals; after all, one cannot *see* radio, television, or radar waves. However, light is, in fact, actually extremely high frequency *radio* waves. Red light has a wavelength of about 600 nanometers (1 nanometer equals 10^{-9} meter), about 130 *million times* shorter than an 80-meter signal. Blue light has a wavelength of about 400 nanometers. The Amateur Radio high frequency bands span from 10 meters to 80 meters, a ratio of 1:8. The visible spectrum extends from about 400 nm to 650 nm, a ratio of 1:1.6.

Most people know that white light is made up of many colors or wavelengths. If we could see radio signals, the myriads of different wavelengths would appear white. But just as radio signals can be separated according to wavelength using narrow filters — i.e., radio receivers — white light can be separated by optical filters. To determine the strength of each wavelength present in a radio band, a spectrum analyzer — generally in the form of a rapidly tunable receiver that scans the wavelength or frequency band and displays its output on an oscilloscope — is used. To determine the wavelength or frequency of a segment of the visible spectrum, an optical spectrum analyzer is used.

Complex, high-resolution optical spectrum analyzers, or spectroscopes, spectrographs, or monochrometers, can be built using film or high gain electronic detectors. But low-resolution optical spectrum analyzers can be built more cheaply and more simply than either radio frequency spectrum analyzers or sophisticated electronic optical analyzers because the human eye can serve as the detector.

The simplest spectroscope is the prism. Newton is known for his work with these: he was the first to apply the word "spectrum" to the band of colors produced by the prism.

Some of Newton's experiments can be duplicated at home. For example, you can demonstrate the separation of white light into colors by building the simple low-resolution system shown in **fig. 1**. First make a small hole (1/16 to 1/8 inch) in a sheet of dark, opaque paper or thin sheet metal. Prop the sheet up against a sunny window, blocking out all light except that which escapes through the hole. Hold a prism about three feet (one meter) from the hole, and a spectrum will appear on the white card as shown. While the colors produced by this system appear to be continuous — red, orange, yellow, green, blue, indigo, violet — dark, narrow gaps can be detected with higher resolution.

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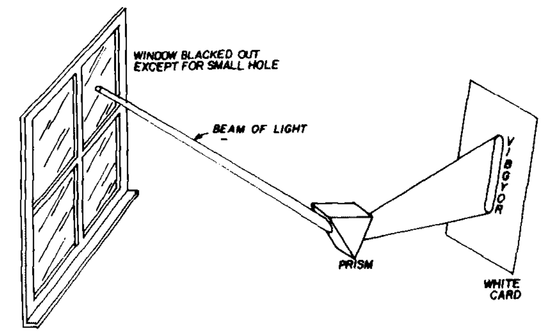

fig. 1. Repeating Newton's experiment with the first optical spectrum analyzer.

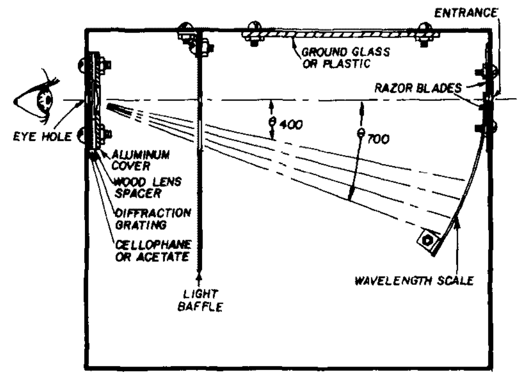

fig. 2. Mechanical illustration of an optical spectrum analyzer.

A higher resolution spectroscope, with a calibrated scale, can be built for less than ten dollars. The design that follows is neither unique nor critical; it works extremely well — considering the cost — and should provide hours of interesting observation. The layout of the spectroscope is shown in **fig. 2**.

design approach

The first thing you may notice is the absence of a prism. A diffraction grating is used instead, because the grating is cheaper than a prism and more easily calibrated. The transmission type grating is made with many finely spaced grooves — 13,400 lines per inch, or 528 per millimeter. (A good microscope would be needed in order to see the grooves.) The grating is available from Edmund Scientific,* in two forms, sheets or slides. The sheets (No. E40267) measure 8-1/2 inches × 11 inches (22 × 28 cm) and sell for $4.95. The slides are small pieces cut from sheets and mounted in 35 mm slide mounts (No. E30282) at $8.50 for 40 slides. The slides are convenient to handle, but I prefer the sheets because of their increased versatility. (A word of caution, however: don't touch, brush, or attempt to clean the delicate grating surface; you'll destroy it.)

construction

Next to the grating is the lens, which should have a focal length equal to or slightly longer than the box you use. The lens serves to collimate, or yield parallel, the light incident on the grating from the entrance slit, which is one focal length or slightly less from the lens. In this position, the slit is imaged to be at or near infinity.

I used a BUD-AU1209 box measuring 6 × 5 × 4 inches (15 × 12.5 × 10 cm); my lens had a focal length of 6 inches (15 cm). The lens should be at least 1/4 inch (1 cm) in diameter. Mine measured 1/2 inch (2 cm). A simple plano-convex, flat-bulged lens works well. (An achromat would be wasted in this application.) Suitable lenses are available from Edmund Scientific. The lens is clamped against the eye hole wall. Then a thin piece of cellophane or acetate is placed against that wall, followed by the grating and two paper spacers. The lens is positioned next, with its curved side facing the slit, a wooden spacer set around the lens, and finally the aluminum plate clamp. (The assembly is held in place with two No. 6 sheet metal screws.)

The eyepiece hole measures 1/8 inch (4 mm) in diameter. A larger hole would produce a brighter image, but parallax with the scale would increase. A horizontal slot 1/8 inch × 1/4 inch (4 × 8 mm) would be best. The spectrum is sighted through the eye hole. *Looking straight through, the zero order or direct* light is seen. Looking downward through the hole, the spectrum will be seen superimposed on the wavelength scale.

The entrance slit is formed with two razor blades. I used single edge industrial blades with the reinforcing splines removed. Each of the blades is held in place with a No. 4 screw, nut, and washer. The slit *must* be horizontal for this design. The slit height (called "slit width," even when the slit is actually horizontal) is a compromise between image brightness and resolution: the wider the slit, the brighter the image. The effective length of the slit is determined by the 1/4 inch entrance hole. Mount the upper blade first. Then, looking through the eye hole, point the spectroscope at the sky or a shaded lamp. Mount and adjust the lower blade to produce a thin, horizontal, even line of light.

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table 1. Values of the angle θ and its sine value throughout the visible light spectrum.

wavelength (nanometers)	θ	$\sin\theta$
400	12.18	0.211
450	13.73	0.237
500	15.30	0.264
550	16.87	0.290
600	18.46	0.317
650	20.06	0.343
700	21.67	0.369

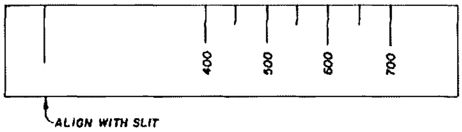


fig. 3. Wavelength scale pattern (not to scale). Actual size will depend on focal length of lens (see text).

Then tighten the lower screw. You can experiment with the slit width after you become accustomed to using the instrument.

calibration

The wavelength scale is easily made and calibrated. Unlike a prism, the dispersion of a grating is given by a simple formula:

$$n\lambda = d \sin\theta \quad (1)$$

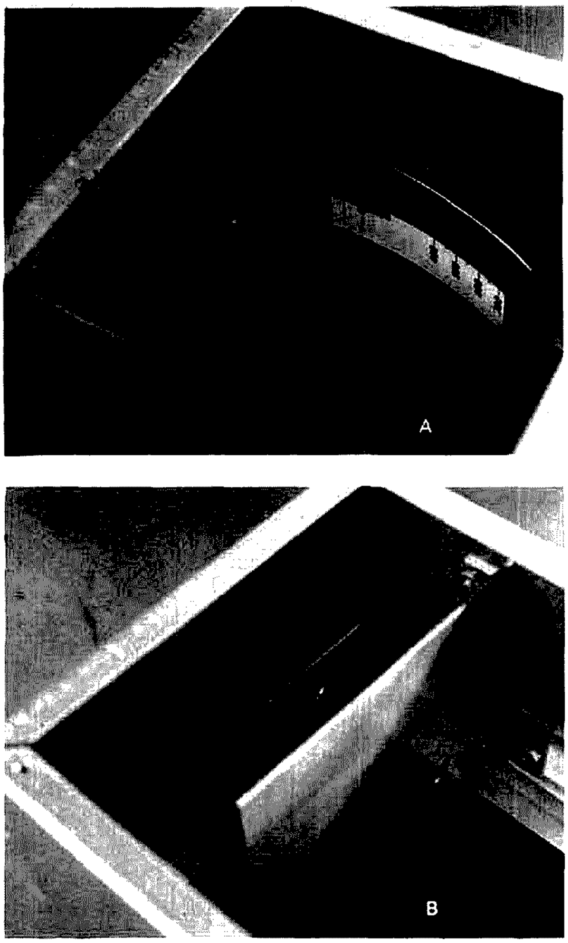
where n is the spectral order, in our case 1, and λ is the incident wavelength. The grating line spacing is d and the angle at which you must look down through the eye hole to see that wavelength is θ.

Rewriting, $\sin\theta = \frac{\lambda}{d} = 0.0005276\,\lambda$ (in nanometers). **Table 1** is a listing of values of θ and $\sin\theta$ for the visible spectrum from 400 nm to 700 nm. The wavelength scale is based on this table. The scale is made on a piece of white paper pasted on an aluminum strip bent into an arc. The center of the arc is the eye hole. The arc radius is the distance from the lens to the scale, about 5.8 inches (14.5 cm) in my model. Laying out the scale is not difficult. First the linear distance along the scale that represents one degree must be determined. I call this number "L."

$$L = \frac{2\pi r}{360} \quad (2)$$

where r is the arc radius.

The scale mark for 400 nm should be 12.18 degrees from the slit position. Mark the slit position, measure out L $\times$ 12.18 inches, and then mark and label the



Interior view *(A)* shows razor blade slit and wavelength scale. Note ground plastic scale light diffuser at left. View *(B)* shows light baffle and lens mounting arrangement.

400 nm line. Similarly multilply each value of θ by L, measure out from the slit position, then mark and label the point. The scale is held in place with an angle bracket. The zero or slit position is aligned with the slit. A rectangular hole on the top of the instrument admits light to illuminate the scale. A ground glass or ground plastic sheet diffuses the illumination to smooth it out and reduce glare. A light baffle prevents the illuminating light from reaching the eye hole directly. The baffle and entire interior is painted flat black. (Obviously, the slit blades, scales, lens, and grating should not be painted.)

Few dimensions are shown in **fig. 2** because your final dimensions will depend on the focal length of your lens and the size of the box. I recommend breadboarding the instrument before final construction.

Once assembled, the spectroscope is ready for final alignment and use. Only two parts should need adjust-

ment: the slit, which should be horizontal and narrow, as previously mentioned, and the grating, which must be rotated so that the grooves are parallel to the slit. This is done by pointing the instrument at a nearby unshaded lamp and rotating the grating until the spectrum is perpendicular to the slit and falls on the wavelength scale.

applications

The spectrum produced by an incandescent lamp is called a continuum because it is continuous from the infrared to the ultraviolet wavelengths; watch what happens when you place cellophane or colored glass filters in front of the slit. Discharge lamps such as neon bulbs and ornamental signs produce line spectra — that is, they emit most of their energy on discrete narrow wavelengths which appear as thin lines in a spectroscope. Try looking at a distant mercury or sodium street lamp with this instrument.

While the sun should never be observed directly (irreversible eye damage may result), the slit width can be reduced greatly and sunlight *reflected* from a distant object — a car, for example — may be observed to produce a continuous spectrum crossed with thin, dark horizontal lines caused by absorption by gases in the solar and earth atmospheres. If the same gases were ionized, bright emission lines would be seen instead.

ham radio

graphic filter design

Response to pulse or steady state signals determines design approach

What would you say if I were to claim that the main burst of sound resulting from hitting a bell with its striker is not "ringing?"

I don't want to hear it.

But that statement would indeed be true if the definition of "ringing" as used by most texts on electronic filter design were applied to a church bell, because what happens when a bell is struck is analogous to what happens when a noise pulse rips into an electronic filter.

It would appear that those who wrote about video pulse amplifiers years ago defined ringing in a peculiar way which then became extended, by word of mouth, to apply to bandpass filters. As a consequence, "flat-top," or Butterworth filters acquired a bad reputation[1]; at the very least, exaggerated benefits for "round top," or Gaussian filters, were implied. Soon the general notion was that flat-top filters ring and round-top filters do not.

In fact, this happens to be true — *but only in the case of low-pass filters driven by a video pulse.*

All band-pass filters "ring," but there are important differences. Assuming equal bandwidths, number of poles, and a unit impulse excitation (much like a sharp noise pulse in our communications systems), a round-top filter will ring with a higher peak amplitude but for a shorter time than a flat-top filter. And a flat-top filter not only rings, but also undulates, which corresponds to what is called "ringing" in lowpass filters. The basic ringing frequency is at or near the filter's center frequency; the undulation frequency is proportional to one-half of the filter bandwidth. In exchange for this ringing difference, the flat-top design provides steeper skirts with slightly more difficult design requirements. **Fig. 1** provides a general picture showing several filter forms with their response to impulse noise and steady-state frequency responses for comparison.

It seems to me that the criteria for whether or not a filter's ringing (or ringing plus undulation) characteristic is a problem should be based upon the real signal environment and how the signals are detected. Clearly a laboratory instrument may well utilize an 8-pole Butterworth with a 10 Hz bandwidth in the metric evaluation of noise-free steady state signals. But that same filter would be useless to those of us who wish to copy CW with our

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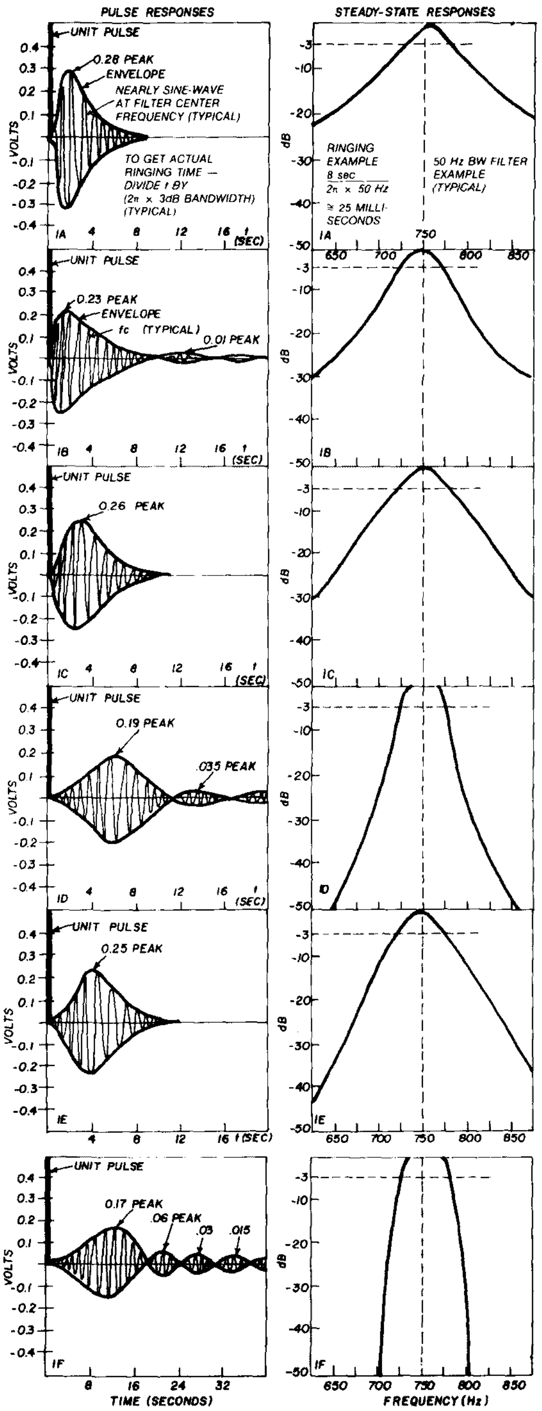


fig. 1. Pulse and steady-state filter response is compared for Gaussian and Butterworth types: *(A)* 2-pole Gaussian; *(B)* 2-pole Butterworth; *(C)* 4-pole Gaussian; *(D)* 4-pole Butterworth; *(E)* 8-pole Gaussian; and *(F)* 8-pole Butterworth.

table 1. Multiplication factor showing stage bandwidth increase necessary when cascading a number of stages to arrive at a required system bandwidth.

number of poles	stage BW multiplication factor
2	1.56
3	1.96
4	2.30
5	2.58
6	2.86
7	3.10
8	3.32

logarithmic hearing response and a signal environment that almost always contains a train of randomly spaced noise pulses both from natural and manmade sources. Because of this, we are forced to apply nearly equal weight to a filter's impulse and steady-state characteristics in our receiver designs.

Fig. 1D illustrates an example using the data on ringing. For a 100 Hz bandwidth design, the main ringing response to a noise pulse would rise and fall in about 13 milliseconds. If you include the time for the second undulation, which is around 16 dB below the main energy bundle, the duration would be 26 milliseconds.

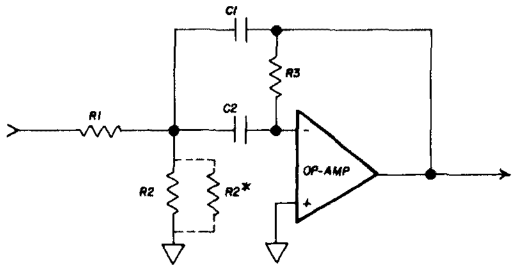
To determine how this may affect CW signal copy, consider that the length of a CW "dit" at 20 WPM is approximately 50 milliseconds. Because in my experience it seems that the human brain is able to ignore energy bursts lasting for fifty pecent or less of the time duration of the basic dit element, this filter appears generally acceptable.

In addition to ringing time, a filter creates a time delay and an increase in rise time (the rate at which the ringing builds up). This provides a secondary benefit in that the clicks and pops of impulse noise and/or key clicks are softened. Consequently, a total rise and fall time of ten milliseconds or more is satisfactory[1] and a lot easier on your ears.

For wide filters such as an 8-pole Butterworth with a nominal 3 kHz bandwidth, the ringing time — even including the third undulation — is only 1.5 milliseconds. While a frequently used IF filter of this type will not confuse CW copy with its ringing, it will not soften those signals that have key clicks.

using multi-pole filters

In general, there are two ways to design multi-pole filters: the ladder, or lattice network method, and the direct synthesis method, in which each pole is isolated. RF and IF filters usually use the first type; audio filters using op-amp active filter elements use the second. In the method using active filters, each op-amp circuit



$Q = (R3)\pi f_c C$

$\text{GAIN} = A = \frac{R3}{2R1}$

$R2 = \frac{Q}{(2Q^2 - A)2\pi f_c C}$

WHERE $C1 = C2 = C$

f_c = CENTER FREQUENCY

$Q = \frac{f_c}{3\,\text{db BW}}$ BW = BANDWIDTH

$R2^*$ IS PAD TO FINE TUNE f_c

fig. 2. Typical bandpass active filter circuit and design equations.

represents one pole (actually a complex conjugate pole-pair, but since our filter designs are derived by transforming lowpass prototypes, the pair is referred to as one pole) and the very low output impedance of the op-amp provides isolation between it and the following stage. Using the basic active filter design of **fig. 2**, just cascade these circuits together with each one tuned to a required frequency and Q to get just about any function you want within the gain-bandwidth limitations of the selected op-amp. Using this basic building block, design examples are given for Gaussian and Butterworth filter types.

Gaussian filters

Round-top filters, usually implemented by the synchronous tuning of a number of stages, are a member of the Gaussian class. Although relatively easy to design, they have one peculiarity that reduces their popularity: as you add poles you must also increase the bandwidth of each stage to obtain a given bandwidth. As a result, ringing time increases *very little* as stages are added, and the filter system skirts do not fall as fast with added stages as they do with the Butterworth class. To design a multipole filter of this type you first decide the system bandwidth you want then determine what the bandwidth of each stage must be from **table 1**, or

$$BW stage = BW system/(2^{1/n} - 1)^{1/2} \qquad \textbf{(1)}$$

where n = number of stages.

For example, if a 4-pole round-top filter with 3 dB points of 700 and 800 Hz is needed, then the design center frequency is $(700 \times 800)^{1/2} = 748.33\ Hz$ and the bandwidth of each stage is $100 \times 2.30 = 230\,Hz$ for a design Q of 3.25. Since each stage is tuned to the same frequency, the gain through the filter is the product of the stage gains. Knowing these requirements, the general equations in **fig. 2** can be used to find component requirements for each of four identical stages.

Butterworth filters

To obtain the flat-top Butterworth response, the center frequency and Q of each stage used is different. However, the geometric pattern used to determine requirements is simple, symmetrical and easy to remember and apply. An example of the initial step in designing a 2-pole filter with a 50 Hz 3 dB bandwidth (725 Hz to 775 Hz) is shown in **fig. 3A**. In this case the semicircle end points represent the 725 Hz and 775 Hz 3 dB points of the final response with the poles located symmetrically and separated by 180 degrees/number of poles. Subordinate semicircles complete the design as illustrated in **fig. 3B**, which enables you to determine the center frequency and bandwidths for both stages as shown. Transient and steady-state responses for this filter are shown in **fig. 1B**. **Fig. 4** shows the pattern for three, four, and eight poles. From this basic pattern, the design for any number of poles follows. Also as the number of poles increases, some of the resulting Qs become quite large, with steeper skirts *and* increased undulation. These patterns are normally plotted on a linear, or undefined, scale in a number of texts. This linear construction scale assumes arithmetical symmetry, which is a reasonable approximation only when the bandwidth is less than 10 percent or so of the center frequency. To show how to handle these designs for larger percentage bandwidths,

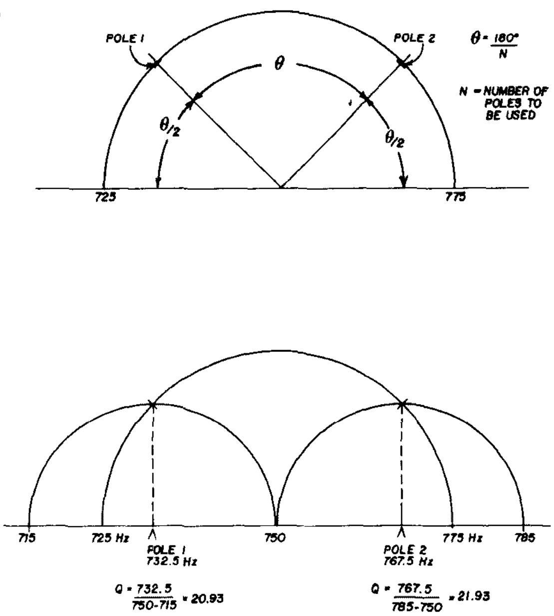


fig. 3. Design sequence to obtain required center frequency and Q for each pole of a maximally flat 2-pole Butterworth filter centered at 750 Hz with 3 dB points at 725 and 775 Hz.

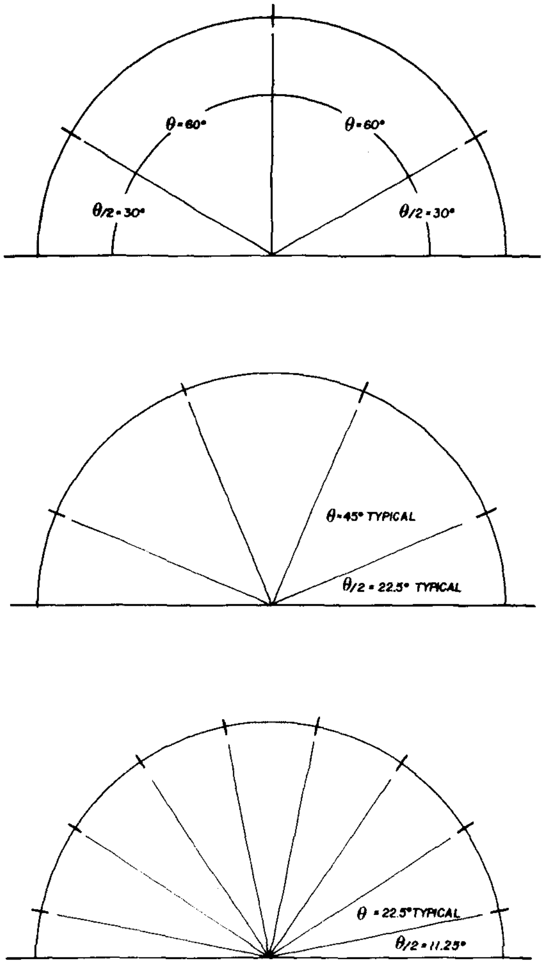


fig. 4. Pole patterns for 3, 4, and 8-pole Butterworth filters.

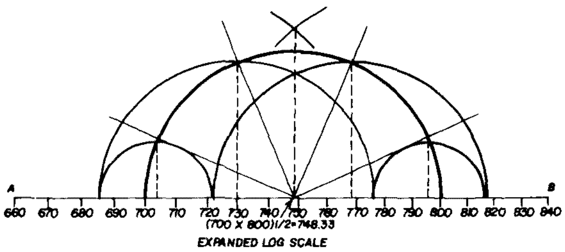


fig. 5. Exact flat 4-pole Butterworth with 3 dB bandwidth from 700 to 800 Hz.

refer to the four-pole example in **fig. 5**. In this figure the only thing that has been changed is the baseline. The physical construction is the same, but by using a log scale the true geometric symmetry of electronic filter circuits becomes apparent resulting in a more accurate design. Clearly, if you design each stage for a gain of 1 there will be some attenuation through this filter type because of the "stagger" tuning. And although it may seem odd, the particular gain of each stage and the order in which they are arranged has no effect on the system frequency response. However, as a practical matter, to avoid a possible reduction in dynamic range, put the low Q stages first with a gain of 1 or so and then design some gain into the latter stages to offset the system loss, if desired. **Fig. 6** shows a complete 4-stage design for both a round-top and a flat-top filter.

active filter component considerations

Using the circuit of **fig. 2** as a reference, the value for R1 is *not* critical, provided it is at least four or five times the resistance of R2. This is usually the case unless you try to obtain a lot of gain with your filters. However, R2, R3, C1 and C2 directly influence the filter stage's center frequency, which is of prime importance. Q, which is the second important parameter, is determined by R3, C1 and C2 – assuming that R2 is adjusted to obtain the design frequency. It is preferable to select the magnitude of C to force R3 to be less than 1 megohm for the required Q to avoid potential problems caused by input currrent of an op-amp. If you intend to build the filter without testing and tuning it, the capacitors should be accurate to ± 1 percent or better, within the temperature range in which most ham equipment is used.

NPO ceramic, mica, or polystyrene capacitors are excellent, but expensive and hard to find. I have had good results for typical ham shack environments using mylar-film capacitors, which are reasonably priced and available in ± 1 percent units.

While carbon composition resistors can be employed, they are difficult to use because they tend to experience semi-permanent value shifts when temperature-cycled by a soldering iron and also over time. Data on carbon-film resistors indicates that they are about ten times-better in this respect; ± 1 percent metal-film resistors are better yet. I use carbon-film, ±5 percent, 1/4 watt resistors and mylar-film capacitors for filters with greater than 50 percent bandwidths and switch to metal-film resistors for narrow multi-pole CW filters.

Integrated circuits such as the 741 op-amp are abundant and inexpensive. There are a multitude of others, all with a nominal open-loop gain-bandwidth product of 1 MHz. With these units, if your active filter stages are

designed to have a gain-bandwidth product of less than 1 kHz and an upper frequency limit of 5 kHz or so, you can just forget the op-amp's characteristics when designing the filter. This is assumed in the equations of **fig. 2**. With more exotic op-amps, of course, you can increase the numbers. In addition, if a resistance with the same magnitude as R3 is included at the op-amp's plus input for bias balance, place a large capacitance from the + terminal to ground to avoid high levels of bias current generated noise voltage across this resistor.

tuning active filters

If you stay within the given op-amp constraints, you can be quite confident of a 10 percent or larger bandwidth design when 1 percent components are used — even without testing. However, if components of this accuracy are not available, you will probably need a signal generator, frequency counter, output indicator and a few components.

Since most active audio filters for CW or SSB will use *Q*s of 5 or less, setting them on center frequency by simply using a signal generator at the input to the filter with a level indicator at the output just isn't practical. The filter will be too broad to enable an accurate adjustment of a frequency tweaking pad on R2. One easy way around this, however, is to take advantage of the fact that the phase slope through the filter is quite sharp and at 180 degrees. Knowing this you can add an op-amp summer as shown in **fig. 7**. With the signal generator set on frequency, and the filter close to its final settings, you adjust the null detector for a minimum indication with the amplitude balance pot, R, then adjust the phase control (frequency) pad across R2 on the filter for a better null. Going back and forth a few times with this technique can set your filter on frequency to an accuracy of ±0.1 percent or so.

To assure this accuracy, it is better to fabricate the filter on one day and tune on the next to let the components

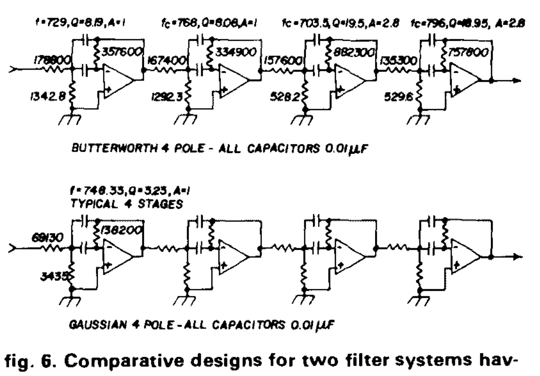


fig. 6. Comparative designs for two filter systems having 3 dB bandwidths from 700 to 800 Hz with unity system gains.

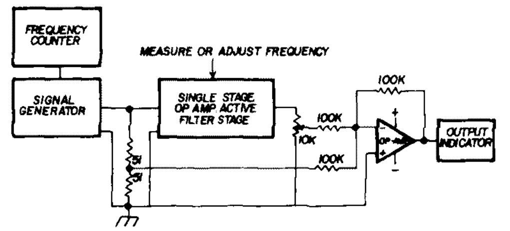


fig. 7. Null-balanced method for accurate determination of filter center frequency.

settle after exposure to soldering temperatures. After you find the required padding resistance, you can then solder in the nearest 5 percent value. Assuming the pad is much larger than the resistor it parallels, it is possible to arrive at 1 percent or better in final frequency adjustment.

Because of the excellent isolation afforded by an op-amp active filter, you just tune each stage independently, then string them together. If you have been accurate with your graph work and calculations, chances are excellent that you will have exactly what you want.

summary

After considering the relative advantages and disadvantages of these two basic filter classes, you probably wish there were some way to get the best of both. I haven't been able to find it, but there is a reasonable compromise that performs well: cascaded Butterworth staggered pairs. Study of the undulation magnitude for the 2-pole Butterworth class shows it to be quite small, which makes the cascading of staggered 2-pole sections attractive. I have used this technique for several years[2] and have found it to be very satisfactory.

Be careful, however, about the apparent simplicity of the synchronously tuned filter type. Lower *Q*s in this case do not mean that you can settle for less accuracy. If you are not careful with this filter, you may wind up with an unplanned stagger — complete with undulation ringing.

references

1. Doug DeMaw, W1FB and Wes Hayward, W7ZOI, "Modern Receivers and Transceivers: What Ails Them?" *QST*, January, 1983, page 11.
2. Don E. Hildreth, W6NRW, "Communications Audio Processor for Reception," *ham radio*, January, 1980, page 71.

Notes:

1. Statements about Butterworth also apply to Chebyshev – *only more so*.
2. PC board and filter parts kit are available from Hildreth Engineering, P.O. Box 60003, Sunnyvale, California 94088. For computer programs listed in Basic for quick and accurate design of filter blocks shown in **fig. 2**, and to calculate expanded log scales as used in **fig. 5**, send $2.00 to cover reproduction, postage, and handling to Hildreth Engineering at the address above.
3. Mouser Electronics at 11433 Woodside Ave., Santee, California 92071, 619-449-2222, is one source of components. (Write or call for catalog.)

ham radio

a switching
high-voltage power supply

Save hours of tedious winding with pre-wound high-voltage coil

The power transformer for my old 5-inch oscilloscope failed recently, leaving me with the problem of how to generate the 800 to 1000 VDC needed for the CRT.

The old transformer used a high-voltage winding with output rectified by a 1V2 tube. My junk box yielded a power transformer that was capable of supplying the heater and low-voltage needs, but had no high-voltage winding. I considered taking it apart and winding on many more turns of No. 36 wire for high voltage, but finally decided there had to be an easier way.

Since switching power supplies are in vogue these days, I designed the circuit in **fig. 1**, using the popular 88 mH toroid coil that is readily available and inexpensive. The beauty of this approach is that all the tedious work — winding the high-voltage coil — is already done. The complete 88 mH coil is used "as-is" for the high-voltage secondary. Two added *windings, the primary and feedback sections*, consist of several dozen turns and take only a few minutes to complete.

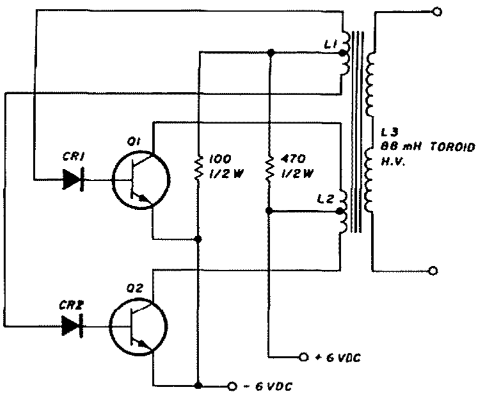
The input current and AC output values shown were measured under no-load conditions. The input current will increase by 100 mA or more under a high-voltage load drawing 0.5 to 1 mA — the usual drain for a CRT. High voltage will also drop several hundred volts under load. I used the voltage-doubler/rectifier circuit shown in **fig. 2**; this yields between 800 and 1000 VDC under load to the 5-inch tube.

the circuit

No particular care need be observed when adding the two windings, other than to keep them in the center of the toroid, away from the ends of the original winding, which will be at a high-voltage potential.

The circuit will work with a wide variety of silicon transistors and I have listed input and output voltages for several varieties. If you have PNPs in your junk box, these can be used by reversing the polarity of the diodes in the base leads as well as reversing the polarity of the DC input. (If you want to use germanium transistors you must change the value of the bias resistors. As a starting point, change the 470-ohm unit to 2000 ohms and eliminate both the 100-ohm resistor and the center-tap connection to the emitters.)

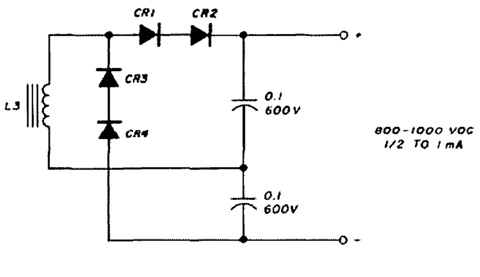
By Jack Najork, W5FG, 3728 East 85th Place, Tulsa, Oklahoma 74136



CR1,CR2 silicon diode, 100 PIV
L1 12 turns No. 28 center-tapped (feedback)
L2 24 turns No. 24 center-tapped (primary)
L3 original 88-mH toroid winding
Q1,Q2 2N3053 or similar
Note: L1 and L2 wound over center of 88-mH toroid

transistor	DC input	no load AC output
2N3394	6 V 160 mA	600
2N3415	6 V 280 mA	750
2N2726	6 V 180 mA	600
2N2222	6 V 280 mA	800
2N3053	6 V 280 mA	800

fig. 1. Switching power supply uses surplus 88-mH toroid with added primary and feedback windings.



CR1, CR2, CR3, CR4: SILICON DIODE, 1000 PIV

fig. 2. Voltage-doubler/rectifier with output suitable for small cathode ray tube.

I used 2N3053s with small, finned heatsinks as switches. These run barely warm under load. 2N2222s will run hot but may be suitable with adequate heatsinks. The 2N2726 and other plastic types would probably be overloaded in this circuit, but, again, may be suitable with good heatsinking. I have not operated the circuit for long periods with these types so I cannot vouch for transistor longevity.

If the circuit does not oscillate, or oscillates with very little AC output when first connected, reverse the feedback winding connections to the diodes in the transistor bases.

ham radio

a programmable
PL tone generator

Produce any frequency from 67 to 250.3 Hz with better than 0.25% accuracy

Now that repeaters have become more common — and their coverage has begun to overlap, especially in metropolitan areas — the use of PL tones has become *increasingly necessary to prevent* bringing up more than one repeater. Because I use a number of repeaters, I wanted a means of generating any PL frequency so that I could bring up any guarded repeater at the flick of a switch.

I knew that the best PL tone generators relied on crystal-controlled oscillators, the oscillator frequency being divided down by some sort of digital counter and then converted into some semblance of a sine wave. This method ensured accuracy and freedom from drift problems. The question was, could a scheme be devised that would produce *all* the PL frequencies with enough precision in a unit of reasonable size and cost?

I did a little arithmetic based on dividing the frequency of a 4.096 MHz crystal I had handy. I found that integer divisors would produce a frequency within 0.5 Hz of the desired frequency in *every* case — and even closer in most cases. Later I changed over to a 3.579 MHz crystal, since this was more readily available. The results were equally satisfactory.

Next came the problem of how to implement the divider scheme. Since I wanted a circuit with a low power drain, I settled on CMOS devices. The *CMOS Cookbook*[1] gave me the two keys I needed: first, a frequency synthesizer IC, the Hughes HCTR0320; second, the digital sine wave generator circuits, which gave me a means of turning any square-wave clock signal into a low-distortion sine wave of lower frequency, with no need for adjustments or tuned circuits.

Basicaly, then, the design is quite simple: a crystal oscillator drives a divide-by-N counter. The output of the counter feeds a digital sine wave generator, which lowers the frequency still further and produces the desired output waveshape.

crystal oscillator

A pair of CMOS gates can form a crystal oscillator with a few passive components.[1] I chose to use NAND gates because this allows a positive-logic enable signal. If pin 1 of U1 (see **fig. 1**) is grounded, the oscillator is disabled. R3 and C3 provide the necessary feedback, and are chosen to give 180 degrees of phase shift at the crystal frequency. The second gate acts as a buffer. The remaining two gates, with U2, are wired as a divide-by-4 circuit. The reason for this will become clear later.

divide-by-N counter

The HCTR0320 consists of the counter, a phase comparator, and a Schmitt Trigger. Because the input waveform has a fast risetime, the Schmitt Trigger is not needed; but its input (pin 16 in **fig. 2**) must be tied to V_{DD} through a pull-up resistor. The phase detector is not used either, so its polarity (pin 21) and reference frequency (pin 18) inputs should be grounded. The oscillator input goes to pin 15; the counter output, taken from pin 14, clocks the sine wave generator.

By Chris Winter, WB0VSZ, 1206 Vincente Drive #F, Sunnyvale, California 94086

The seven binary-mode control lines are typically used in frequency synthesizers to offset transmit and receive frequencies, as for repeater access. They are not used in this design, and should be grounded. The remaining twelve control inputs are set up as three BCD digits. They allow selection of any division ratio between 3 and 999. The required ratios fall between 179 and 668. (See **table 1** for complete information.) Note that, in this counter, the output duty cycle is the reciprocal of the division ratio; the pulses will be very narrow.

digital sine wave generators

The secret of a digital sine wave generator is the Johnson, or walking-ring, counter, in which a fixed pattern circulates constantly. In this case, the pattern is simply a square wave at any output. The various outputs, each at either V_{DD} or ground, are summed through resistors of different values. With proper choice of values, the result is a waveform that closely resembles a sine wave. For a counter of M stages, the clock frequency must be 2M times the desired output frequency, and the lowest harmonic present in the output is of order M-1. This means that a simple lowpass filter can produce a very clean sine wave.

Digital sine wave generators can be made from chains of D-type flipflops like the MC14013. For example, a pair of 14013s makes a 4-stage counter. It requires three resistors and produces a sine wave at one-eighth the clock frequency. This wave will be riding on a DC level of $V_{DD}/2$. The more stages in the counter, the closer the

table 1. PL tone frequencies accurate to within 0.25 percent of EIA standard.

EIA standard PL frequency Hz	value of N	output frequency	error, Hz	error, %
67.0	668	66.98	−0.02	−0.03
69.3	646	69.26	−0.04	−0.06
71.9	622	71.94	+0.04	+0.06
74.4	601	74.45	+0.05	+0.07
77.0	581	77.01	+0.01	+0.01
79.7	561	79.76	+0.06	+0.08
83.5	536	83.48	−0.02	−0.02
85.4	524	85.39	−0.01	−0.01
88.5	506	88.43	−0.07	−0.08
91.5	489	91.50	0.00	0.00
94.8	472	94.80	0.00	0.00
100.0	447	100.10	+0.10	+0.10
103.5	432	103.57	+0.07	+0.07
107.2	417	107.30	+0.10	+0.09
110.9	403	111.03	+0.13	+0.12
114.8	390	114.73	−0.07	−0.06
118.8	377	118.69	−0.11	−0.09
123.0	364	122.92	−0.08	−0.07
127.3	351	127.48	+0.18	+0.14
131.8	339	131.99	+0.19	+0.14
136.5	328	136.42	−0.08	−0.06
141.3	317	141.15	−0.15	−0.11
146.2	306	146.22	+0.02	+0.01
151.4	296	151.16	−0.24	−0.16
156.7	286	156.45	−0.25	−0.16
162.2	276	162.12	−0.08	−0.05
167.9	266	168.21	+0.31	+0.18
173.8	257	174.10	+0.30	+0.17
179.9	249	179.70	−0.20	−0.11
186.2	240	186.43	+0.23	+0.12
192.8	232	192.86	+0.06	+0.03
202.7	221	202.46	−0.24	−0.12
203.5	220	203.38	−0.12	−0.06
210.7	212	211.06	+0.36	+0.17
218.1	205	218.26	+0.16	+0.07
225.7	198	225.98	+0.28	+0.12
233.6	192	233.04	−0.56	−0.24
241.8	185	241.86	+0.06	+0.02
250.3	179	249.97	−0.33	−0.13

output is to a true sine wave. The MC14018 is very convenient here because it has five suitable stages in one package. A pair of 14018s is doubly convenient because a 10-stage counter divides the clock frequency by 20.

Only an additional factor of 4 is needed to bring the total divisor within range of the HCTR0320. This factor, of course, is supplied by U2.

The resistor values shown in the sine wave generator

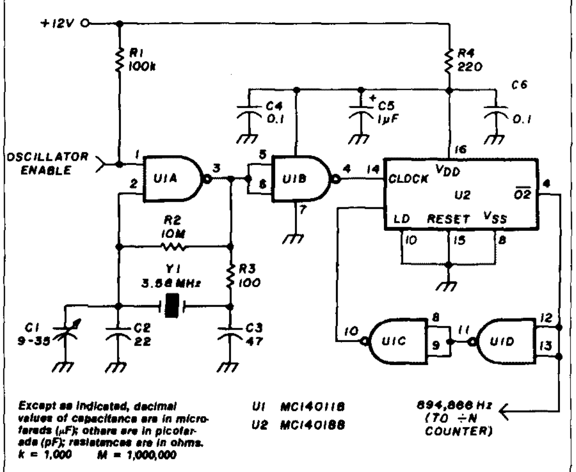


fig. 1. Clock frequency for the PL tone generator is produced by a crystal oscillator and divide-by-four circuit.

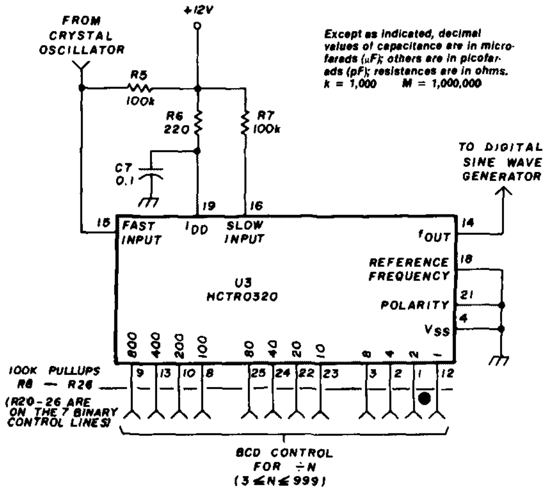


fig. 2. A Hughes frequency-synthesizer IC divides the clock frequency by the desired integer.

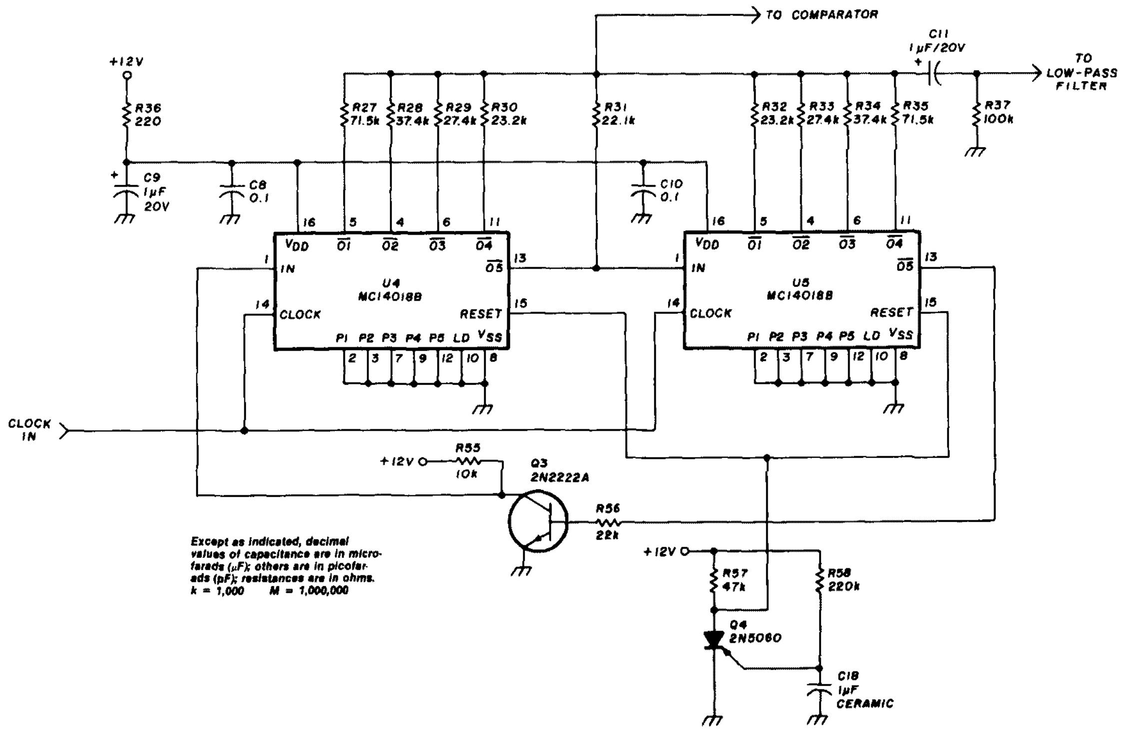


fig. 3. A 10-stage digital sinewave generator produces an output with very low harmonic distortion.

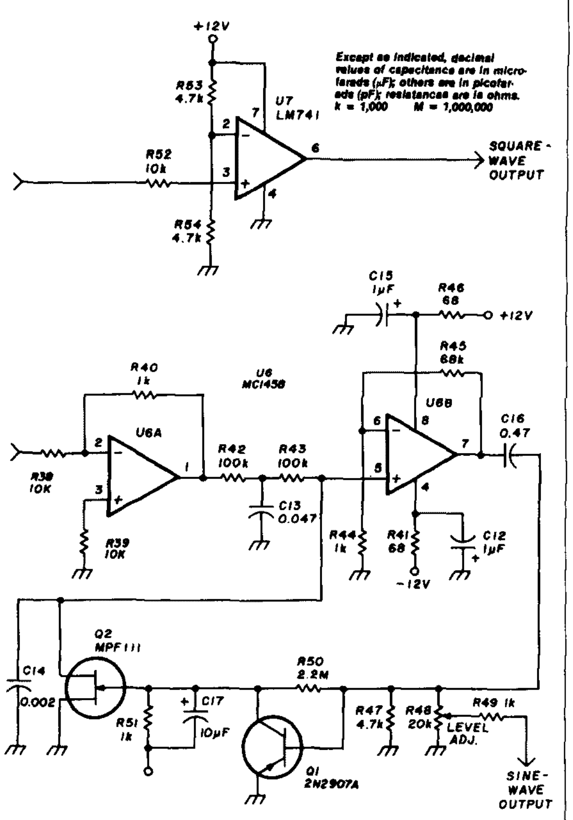


fig. 4. A lowpass filter and AGC amplifier condition the sinewave output. U7 provides a test signal for frequency checks.

parts list

item	*description*
C1	9-35 pF trimmer
C2	22 pF 100V 5 percent mica
C3	47 pF 100V 5 percent mica
C4,C6,C7, C8,C10,C19	0.1 μF 50 ceramic
C5,C9,C11, C12,C15	1 μF 20V tantalum
C13	0.047 μF 50V ceramic
C14	0.002 μF 50V ceramic
C16	0.47 μF 50V ceramic
C17	10 μF 20V tantalum
C18	1 μF 50V ceramic
Q1	2N2907A
Q2	MPF111
Q3	2N3704
Q4	2N5060
R1,R5,R8-R26, R37,R42,R43	100k 1/4 watt 10 percent
R2	10 meg 1/4 watt 10 percent
R3	100 ohm 1/4 watt 10 percent
R4,R6,R36	220 ohm 1/4 watt 10 percent
R27,R35	71.5k 1/8 watt 1 percent
R28,R34	37.4k 1/8 watt 1 percent
R29,R33	27.4k 1/8 watt 1 percent
R30,R32	23.2k 1/8 watt 1 percent
R31	22.1k 1/8 watt 1 percent
R38,R39,R52,R55	10k 1/4 watt 10 percent
R40,R44,R49,R51,(1 MHz)	1k 1/4 watt 10 percent
R41,R46	68 ohm 1/4 watt 10 percent
R45	68k 1/4 watt 10 percent
R47,R53,R54	4.7k 1/4 watt 10 percent
R48	20k potentiometer
R50	2.2 meg 1/4 watt 10 percent
R56	22k 1/4 watt 10 percent
R57	47k 1/4 watt 10 percent
R58	220k 1/4 watt 10 percent
U1	MC14011B
U2,U4,U5	MC14018B
U3	HCTR0320
U6	MC1458
U7	LM741
Y1	3.579545 MHz crystal

PCB1 is a 3 × 5 inch (7.62 × 12.7 cm) double-sided PC board with plated-through holes. It mates to double-readout 15-position edge connector. A bare board with complete documentation is available for $15 from: Zombico, P.O. Box 61831, Sunnyvale, California 94088.

circuit (**fig. 3**) work only for a 10-stage counter. (The *CMOS Cookbook* includes information on values for counters of various lengths.) Precision resistors are nice, but quite acceptable results can be obtained with ordinary carbon 10 percent tolerance resistors.

Two pitfalls should be mentioned. The first is the fact that at power-on, the 14018s may contain an undesired pattern of bits; in a sine wave generator, this will produce really incredible waveshapes. The solution is to include a power-on reset circuit. The other pitfall is the need for an additional inverter between the two 14018s. While the *Cookbook* does not emphasize this point, an inverter is necessary to obtain the proper waveshape and division ratio. In my design, there was no room for another IC so the SCR does power-on reset, and Q3 provides logic inversion.

other features

I included some signal conditioning in the design. There is a lowpass filter, of course, and a blocking capacitor to remove the sine wave from the half-V_{DD} DC level. Also included are an output level adjustment, and a circuit to hold the level constant.[2] For ease in checking the operation of the unit with a frequency counter, a comparator provides a square wave at the output (PL) frequency. These features are shown in **fig. 4**.

application information

The unit is simple to use. Voltages required are +12 and −12 volts at a current drain of about 20 mA. Maximum output is better than 1 V_{p-p}. The divider control lines can be wired to three BCD thumbwheel switches; a 12-bit latch controlled by a computer could be used instead. The PCB layouts are shown in **fig. 5** on page 56.

references

1. Don Lancaster, *CMOS Cookbook*, Howard W. Sams Co., Indianapolis, Indiana, 1977. (Available from Ham Radio's Bookstore, Greenville, NH 03048, $14.95 post paid.)

2. Chris Winter, WB0VSZ, "Tone Encoder for 2-Meter Autopatches," *ham radio*, June, 1980, page 51.

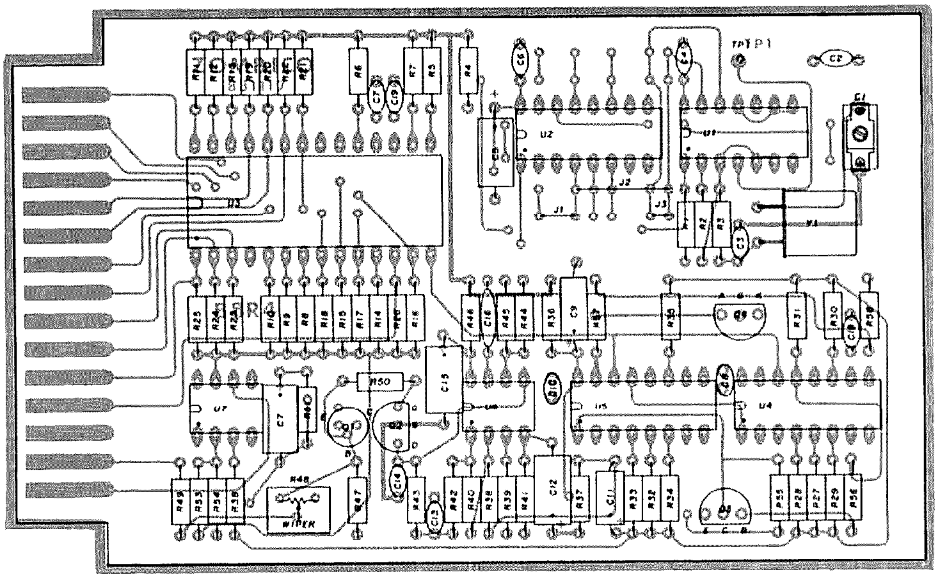


fig. 5A. PL tone generator, PCB component side.

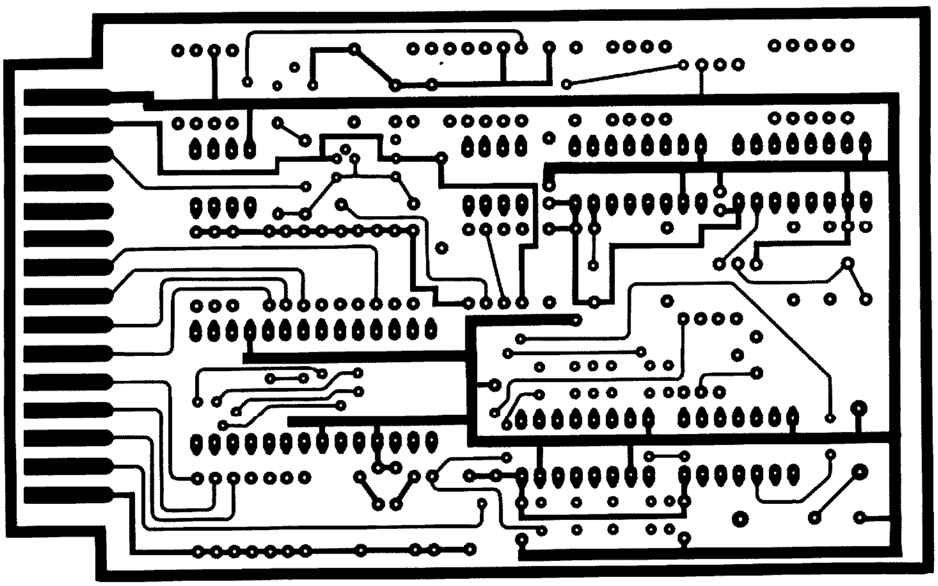
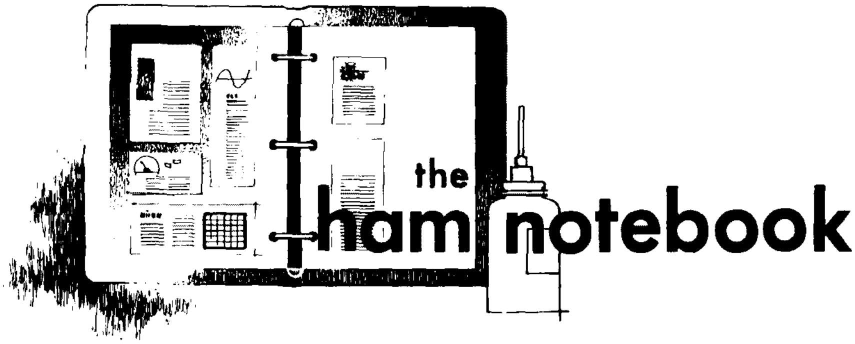


fig. 5B. PL tone generator, PCB ground plane side.



Ten-Tec Corsair modification

Separation of the AGC and RF/IF gain functions in modern transceivers provides improved receiver operation. In a receiver in which the AGC remains on when the RF/IF manual gain control is backed off from the maximum position, SSB reception can be improved by adjusting the gain until the received noise signal is just above the receiver minimum-signal threshold. The Ten-Tec Corsair incorporates separate receiver AGC and manual IF-gain controls.

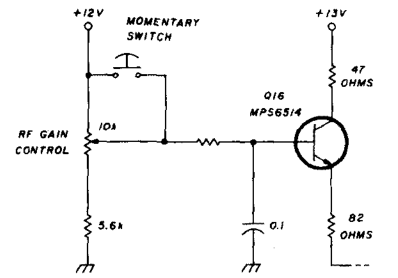


fig. 1. The addition of a momentary switch provides normal operation and instantaneous S-meter check capability to the Ten-Tec Corsair.

When operating the Corsair with the IF-gain turned down, the S-meter readings for a given signal strength are reduced. In order to obtain meaningful S-meter readings, it is necessary to momentarily turn the IF gain to maximum. This modification provides that function through the addition of a "momentary" switch as shown in **fig. 1**.

The switch is a subminiature, normally-open, spring-return push button. To take an S-meter reading the button is depressed shorting the wiper to the top of the control. Due to the excellent AGC characteristics of the receiver, there's no blast from the speaker.

There's a convenient space on the Corsair's front panel for mounting the push button — to the right and above the center of the manual IF-gain control (*RF GAIN*). Radio Shack subminiature push-button switch No. 275-1571 is a good, if not exact, match for the SPOT push button on the Corsair panel.

This technique can probably be adapted to other transceivers with an AGC function that remains operative over the full range of the RF/IF manual gain control.

Jack Geist, N3BEK

visual telephone ring indicator

At the company where I work, there are four offices — each with its own telephone — along a corridor, with a conference area across the hall. Whenever the staff was gathered in the conference area and one of the office phones would ring, it was impossible to tell whose phone it was; obviously, a light above each office door to indicate a ringing telephone would minimize interruptions and save steps.

The telephone company offered to install such a device for $52 per phone, and charge a monthly rental fee as well. **Fig. 1** shows how I solved the problem inexpensively and without the phone company; perhaps it will be of

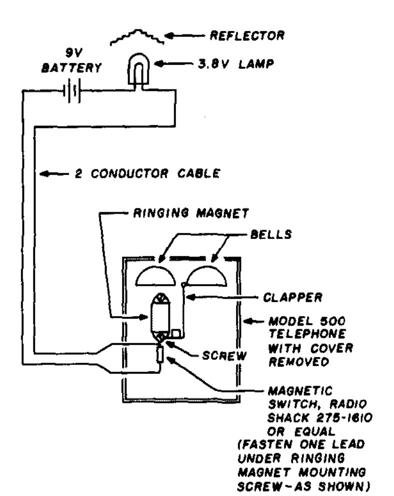


fig. 1. Circuit for simple telephone-ring indicator.

special interest to deaf persons and to all who use RTTY to communicate with them.

J. W. Dates, W2QLI

quick fix for soldering irons

Tired of paying high prices for soldering gun tips? Make your own out of No. 10, 12, or 14 copper wire. Just tap the two existing holes to take a No. 8 bolt and insert your new tip.

There's also no need to pay five or six dollars for a new switch. When a switch on my Weller iron stopped working, I drove out the pin fasteners with a small nail and carefully removed the cover. Sure enough, I found a broken spring. I found a replacement spring at the hardware store for forty cents and cut it in half. I then squared off the cut end, holding the switch assembly with my thumb, and worked the new spring into the case. Putting on the cover and replacing the pins, I wired in the switch and reassembled the gun.

John Labaj, W2YW

portable shortwave receiver

Compact unit covers 46 through 100-meter bands

The design of this compact, high performance shortwave receiver has evolved over several years.[1,2,3] Its underlying concept, namely low power consumption, makes it ideal for DXpeditions. The unit provides spurious-free, stable operation and sensitivity comparable to commercially-manufactured receivers.

While the earlier projects sported a digital frequency readout using a mechanical counter with a 455 kHz IF strip and mechanical filter,[4] the current design includes electronic frequency readout and higher frequency IF (in the 9-10 MHz range). With the primary objective of low power drain, a multiplexed counter was used, incorporating the Intersil 7207-7208 IC combination. In this configuration a frequency response exceeding 25 MHz was achieved.

To avoid adding additional ICs to this project, the VFO output was read directly without any signal conditioning. For receive it was desired that the displayed frequency be equal to the tuned frequency; there are several methods of accomplishing this.

Use of a 10 MHz IF strip and a VFO tuned above the IF frequency is one method. By deleting the first digit of the display, one reads the true receiver frequency.[4] Use of a 9 MHz IF strip and a VFO tuned above the HF frequency is another method. Again, deleting the first digit from the display will produce a readout indicating exactly 1.0 MHz above the received frequency. By rewiring the MHz digit, this can be made to display 1.0 MHz lower (i.e., than the true tuned frequency).

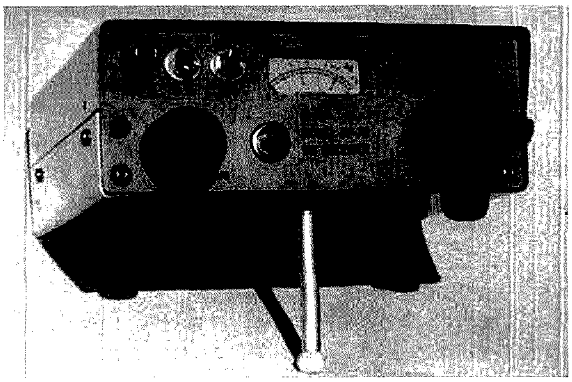


fig. 1. Front view of receiver shows unit mounted on 4-inch removable slip-on tilting pin, which eliminates parallax effects on S-meter readings and frequency display. Panel controls (left to right) include: antenna, RF bandswitching, VFO bandswitching, and gain knob (below). S-meter is in center, tuning knob is at right, just below 4-digit display.

The first method is particularly convenient from a mechanical standpoint. Judging from the response to earlier articles, most builders have had difficulty in locating a convenient source of 10 MHz crystal filters. Considering the costs involved, it may pay to build your own crystal filter.[5,6]

Other methods (still using a 9 MHz filter and a VFO tuned above IF) involve reading only the last three digits (1, 10, and 100 kHz) and reading the MHz number off the skirt of the bandswitch knob. While this may sound a bit crude, it offers the advantage (over bandswitching the MHz digit with diodes) of reduced current drain. It is also possible to control the MHz digit by a push-button switch, and to switch it on only when the MHz reading is actually needed.

One improvement over the earlier designs is the replacement of the antenna trap, set at the IF frequency, by a notch that attenuates IF signal feed-through considerably; this is particularly helpful because in some areas the 10 MHz signal from standard frequency stations may reach a local field strength level high enough to cause performance degradation. Extensive shielding and care in front-end design should be used in such cases.

design

This approach employs the later method and covers 3.0-6.5 MHz in two bands. Its coverage can be expanded to include the 7 MHz ham band or other frequencies of interest by making provisions for proper readout options, and adding more digits, if needed. However, if one prefers to leave this project in its simpler form, satisfactory results can be achieved by adding an external converter.[7]

All bandswitching is performed electronically, using transistors and miniature toggle switches. Because transistors are back-to-back diodes, using input-output signal isolation appears to be better than using diodes. This is particularly true for the VFO circuit.

The power supply circuit takes either 12 VAC or DC at its input jacks and acts as a polarity inversion protector. To keep power consumption down, the counter can be turned off when not in use — for example, when you remain on the same frequency for an extended period of time.

Another improvement is the adoption of a passive balanced mixer, using Germanium diodes; the receiver front-end response thus becomes considerably "cleaner," as cross modulation and spurious response is reduced.

By Jack Perolo, PY2PE1C, P.O. Box 2390, Sao Paulo, Brazil

Driving a passive mixer requires a high injected VFO signal level. Consequently a suitable amplifier was designed for this purpose. Under load conditions, the VFO output is 3.0-3.2 volts. The value of the buffer emitter capacitor was selected for flattest output voltage within the VFO frequency range desired.

construction

The photos (**figs. 1**, **2**, and **3**) illustrate construction details, while the schematics in **figs. 4**, **5**, and **6** provide the electrical circuits.

This set measures 5.5 × 2 × 6 inches (140 × 50 × 150 mm) and weighs 4.6 pounds (2.1 kg), excluding the wall adapter. In Brazil, the cost of parts comes to about $280; the unit requires about 300 hours to build.

Cabinet material is 18 gauge stainless steel, with a 16-gauge front panel to permit better engraving. The back panel and main chassis are made from 1/8 inch thick aluminum; the shields are also aluminum, but only 1/16 inch thick. All PC boards are fastened with 4-40 stainless steel screws, and use 1/4 inch aluminum round pillars.

The weight could be reduced by using lighter components and thinner plates. However, for greater stability, and as a matter of habit, I prefer to use the thicknesses specified.

Some substitution of components may be necessary if specified parts are difficult to obtain. For example, the Polar 104 pF main variable capacitor is a high quality British import, fully silvered, on ceramic insulation, with bearings at both ends. (J.W. Miller used to distribute these in the U.S.) The 10:1 vernier, manufactured by Jackson in the U.K., is also heavy-duty. The 9 MHz crystal filter is Japanese, with a 1.8 kHz bandwidth at −6 dB and 3.5 kHz at −60 dB. (It's available through Yaesu dealers.) Should substitution be required, the German import by KVG is also suitable, and available. The S-meter, of Japanese manufacture, uses a 1 mA movement and is directly calibrated in S-units. The IF coil forms are made in Brazil; their base measures about 0.6 × 0.6 inch (15 × 15 mm), 0.8 inch (20 mm) high, with a coil diameter of 1/4 inch. (J.W. Miller distributes a similar unit in the United States.)

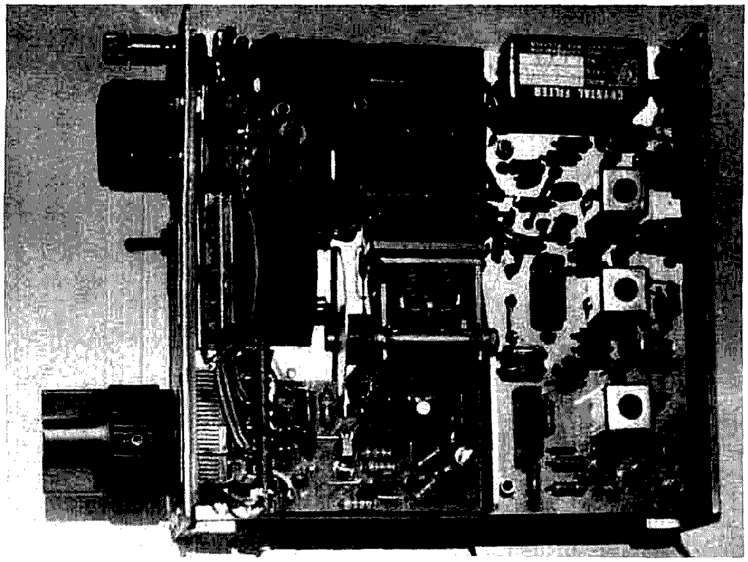


fig. 2. Top view of receiver. *Top left PC* board includes regulated power supply and audio stages. *Right side PC* includes IF stages, detectors, AVC amplifier and S-meter zero setting trimpot on this double-sided board. *Bottom left PC* includes VFO transistor, amplifier, and buffer together with electronic bandswitching components. (Note the use of silver mica capacitors on VFO input circuit.)

Note: The 9 MHz antenna trap is at the top left, near the antenna input. The 9 MHz crystal filter is at the top right, near input of IF PC board. Even though it is not visible in this picture, the crystal filter input pin is shielded below the receiver chassis to avoid stray coupling with the RF/mixer PC, located on the opposite side of the chassis. At bottom left, maintained in a vertical position by aluminum brackets, is a small PC board holding the four-digit display. The back panel contains 12V input jacks, and two phone jacks, the second used for tape recording.

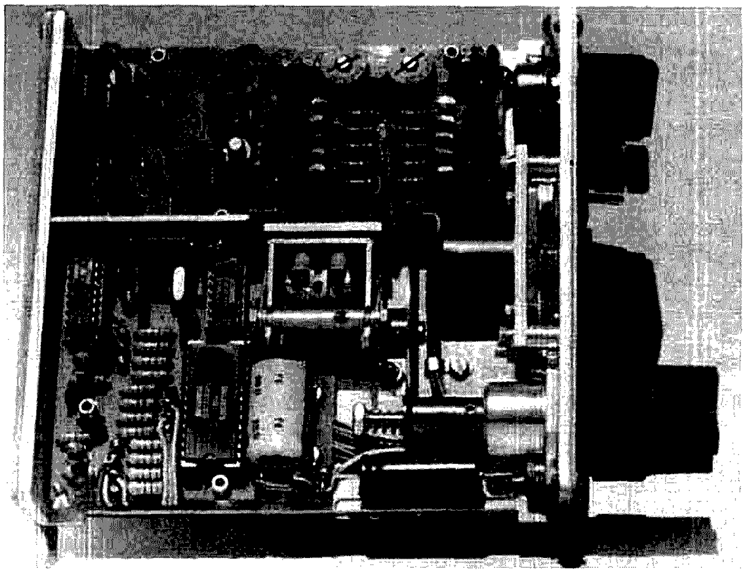


fig. 3. Bottom view of receiver. Top PC: mixer stage is shown on the left. RF circuits, with the toroidal coils and trimmers, are clearly visible. The ten aligned resistors and nearby transistors provide electronic bandswitching for the front end. Note that the 40841 transistor is plugged into a socket, for replacement in case of burnout during field-day operation. Bottom (left) PC houses the entire frequency counter, complete with input signal processing. The tuning (VFO) variable capacitor is in the center, linked by an aluminum disc to the pinch-rim drive coupled to the tuning vernier. Two electrolytic power supply capacitors are visible below the variable capacitor. Note the shielding between the RF/mixer stages and counter, and the bracket on back of the S-meter, holding it in place.

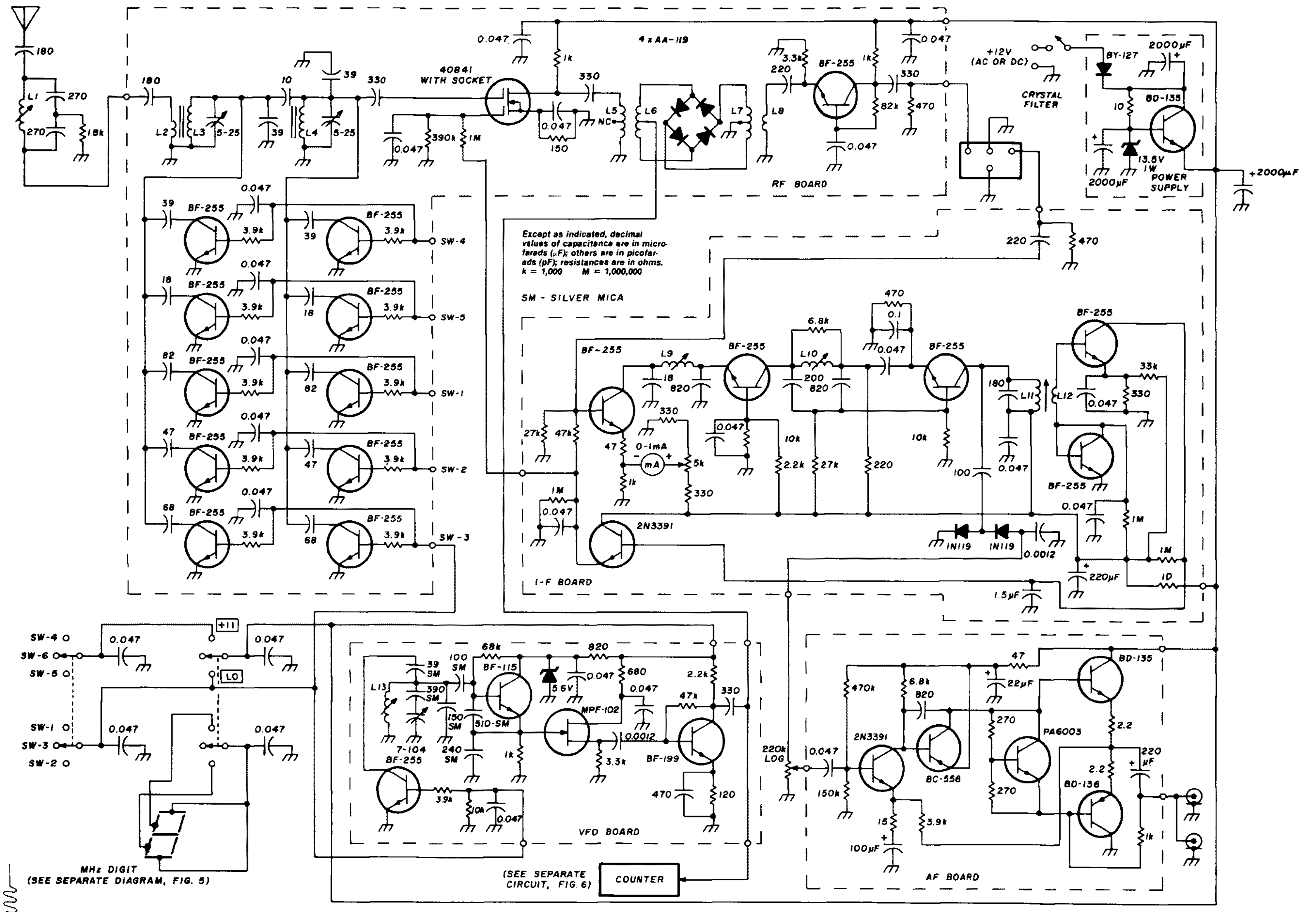


fig. 4. Schematic of portable shortwave receiver.

table 1. Coils required for construction of portable shortwave receiver. (Unless otherwise noted, all wire is No. 28 enamel.)

coil	description	frequency range, MHz
L1	18 turns in 3 layers on a 3/16" slug-tuned ceramic form (No. 31 wire)	9.0
L2, L3	RF coil 6 turns and 50 turns (No. 31 wire), respectively on Amidon T-37-2 toroid	3.0-6.5
L4	RF coil 50 turns No. 31 wire on Amidon T-37-2 toroid	3.0-6.5
L5, L6	mixer input — Quadrifilar wound on a 3/8 × 3/8 × 1/4-inch ferrite bead, 6 turns + 6 turns, with the center tap of the primary coil not connected	
L7, L8	mixer output — Trifilar wound on a 3/8 × 3/8 × 1/4-inch ferrite bead, 6 turns + 6 turns	
L9	first IF coil — 40 turns in 2 layers on a 1/4 inch slug-tuned nylon form with aluminum shield	9.0
L10	second IF coil — 18 turns on a 1/4 inch slug-tuned nylon form with aluminum shield	9.0
L11, L12	third IF coil — 18 turns and 6 turns on a 1/4 inch slug-tuned nylon form with aluminum shield	9.0
L13	VFO coil — 10 turns in 2 layers on a 3/16 inch slug-tuned ceramic form	12.0-15.5

The pinch-rim drive mechanism has an 8:1 reduction ratio and is homebuilt, using a 1/32 inch thick duralumin-um disc and a 1/2 inch brass shaft. This gives the tuning unit a "velvety" feel and, considering that the variable capacitor swing required a rotation of about 160 degrees, an overall reduction ratio of about 1:35.

Depending on the size of the tuning knob, comfortable tuning can be achieved up to a tuning rate of 40-50 kHz/turn; therefore, each bandspread should not exceed (35)(50) = 1,750 kHz.

choosing coils

Table 1 lists the coils necessary for the coverage given; *all* wire is No. 28 enamel, except where noted.

If other band coverage is desired, coils L2, L3, L4 and L13 should be modified accordingly to tune the new bands of interest; all other coils remain unchanged. For the sake of miniaturization, the trimmer capacitor used to adjust the counter readout has been replaced by a fixed capacitor, marked with an asterisk on the schematics. This capacitor should be adjusted for the crystal and the filter center frequency used, generally falling in the range of 15-47 pF. Because this circuit is extremely stable, I haven't noticed any need to recalibrate this unit even after years of operation.

circuit boards

The full project uses six 1/16-inch thick epoxy printed circuit boards as follows:

1. RF front-end and mixer, complete with band-switching, 2.3 × 4.7 inches (58 × 118 mm).

2. 9 MHz IF strip and AVC amplifier, on double-sided board, 2.3 × 4.4 inches (56 × 110 mm).

3. Audio strip and power supply, 1.9 × 2.0 inches (48 × 50 mm).

4. Frequency counter, 2.2 × 2.7 inches (55 × 67 mm).

5. Four-digit display, 0.9 × 1.8 inches (23 × 44 mm).

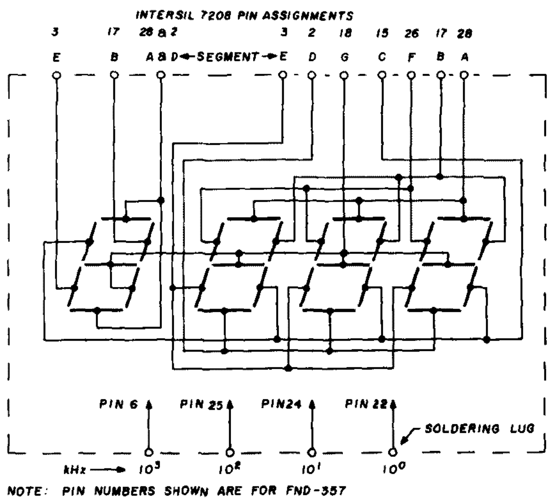


fig. 5. Four-digit display schematic.

6. VFO, amplifier, and buffer, complete with band-switching, 1.4 × 2.2 inches (36 × 55 mm).

These circuits, shown inside the dotted lines on the main schematic (**fig. 4**), were prepared using a resist ink, and then etched in a ferric chloride solution. The double-sided IF PC board is protected with Contac™ during etching. After etching, the PC boards are cleaned with steel wool and painted with acrylic lacquer.

semiconductor devices

Philips makes the BF-115, BF-255, A-119, and BY-126. Siemens supplies the BC-548, BD-135, and BD-136.

The MPF-102 and zeners are from Motorola; the 40841 is made by RCA while the 2N3391 is from Texas Instruments. Substitutions can be made, provided care is taken to confirm equivalency.

ICs 7207 and 7208 are manufactured by Intersil; the 74LS90, a low-drain variety of the 7490, is available from many sources.

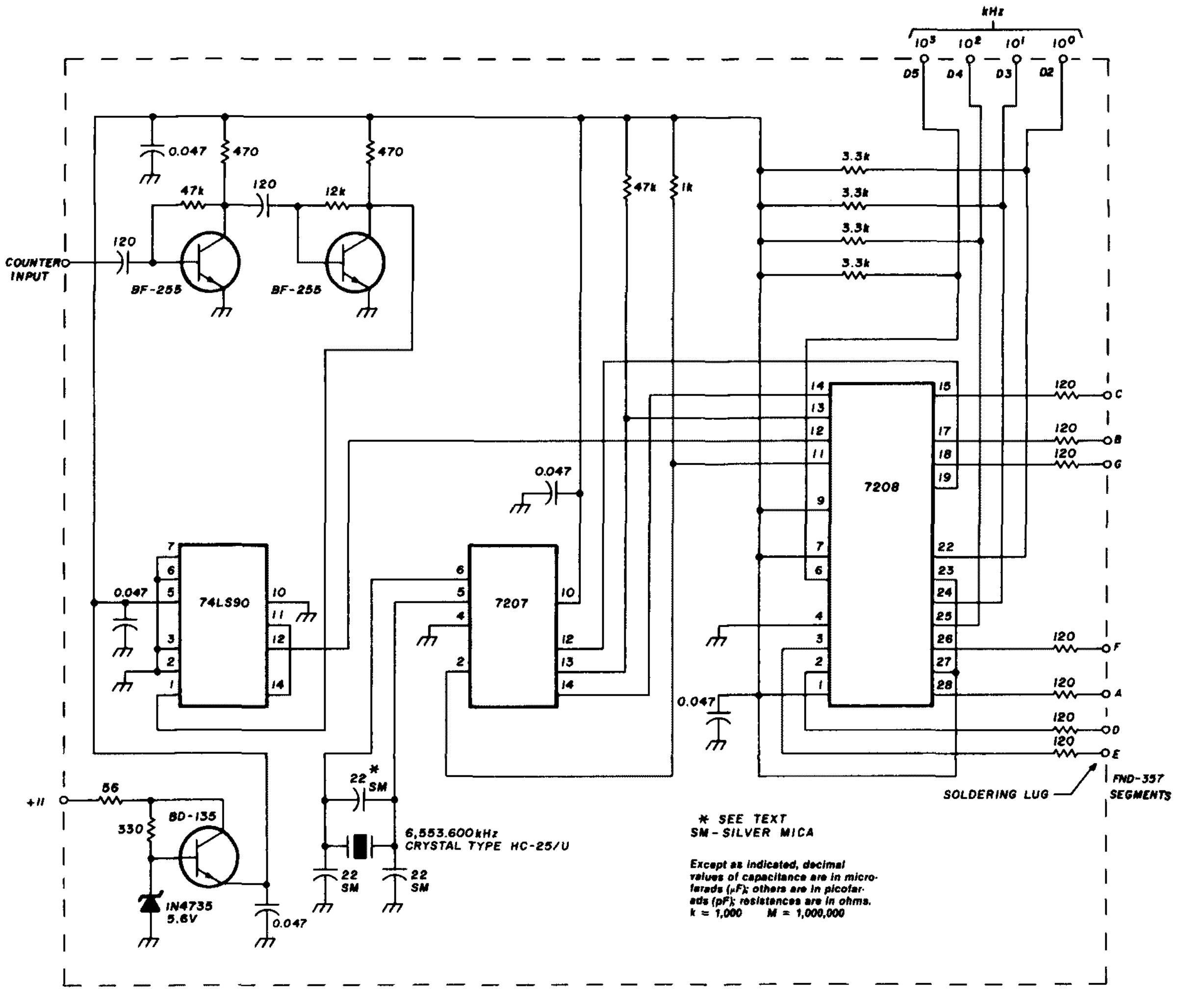


fig. 6. Frequency counter schematic.

The FND-357 display is manufactured by Fairchild.

performance

Current drain on a 12 VDC supply with the display on and the volume set at a comfortable listening level is 90-95 mA; using power supplied by a car battery. Field-day type operation can be extended for several days in a row without any need to recharge. The counter is responsible for 50-55 mA of the current drain. Overall sensitivity, at 5.0 MHz, is about −125 dBm, using the specified crystal filter; loss of sensitivity becomes noticeable when the feed voltage drops below 10 volts.

After a five-minute warm-up period, stability is ±20 Hz/hour at 4.0 MHz (i.e., with the VFO running at 14.0 MHz); this stability is achieved by operating the VFO transistor at low voltage (5.6 volts), by selecting the best capacitive feedback ratio, by using high parallel capacitance and, finally, by installing silver mica capacitors. The counter is sensitive and stable at an input level of 12 mV into 50 ohms.

acknowledgement

I thank Fernando, PY2DQU, for his considerable help on the theory and measurement equipment.

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ham radio

ham radio TECHNIQUES

Bill Orr, W6SAI

a *very* wideband no-tune antenna

It's always refreshing to find an example of original thinking, especially when it pertains to antennas! (The dipole and the ground plane have been reinvented too many times.) This novel antenna was recently described in *Amateur Radio*, the publication of the Wireless Institute of Australia,* a magazine I highly recommend to the readers of this column.

Some months ago I discussed the so-called Australian dipole, a cage-like antenna that covers most of the high-frequency spectrum.[1] This large, heavy array has been widely used in Australia and Africa, but is still virtually unknown in the United States.

Recently, experiments conducted by VK6IM and VK6YX of Western Australia have resulted in a grounded version of the wideband dipole that can be used as a vertical or a sloper antenna. This new design covers the frequency range of 1.8 to 14.34 MHz, with a measured SWR on the feedline across the span of about 1.2:1. Thus it seems the "VK6-broadband" antenna is an excellent choice for multiband coverage with the new solid-state transceivers that demand a low value of SWR on the antenna system.

The schematic of the antenna is shown in **fig. 1**. Basically it is a "fat" monopole with an impedance matching network placed about a third of the way down from the free end. The antenna is worked against a counterpoise ground system and is fed through a simple broadband transformer.

antenna assembly

The "VK-6 broadband antenna" is about 75 feet long and is formed of a five-wire cage about 6 feet in diameter. It is composed of two parts joined by a parallel-connected resistor and inductor. The lower cage section is connected to the feedline by means of a transformer. The upper section is about half the length of the lower section. The five-wire cages are made of No. 18 semi-hard drawn copper wire and the spreaders are 1/4-inch diameter fiberglass rods, each 73 inches long. The rods are drilled in the center and at 1-1/2 inches from each end with a 1/16-inch diameter drill, the end holes being at right angles to the center hole.

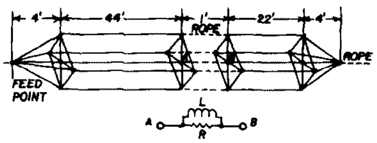

fig. 1. Oblique view of the VK-6 broadband cage antenna. Spreaders are 72 inches long. All five wires are bonded together at each spreader as shown. Connecting ropes join the antenna sections. The matching network is placed between points A and B of the cages. The antenna is fed at the left and a balun is used.

To fasten a cage wire to the end of a spreader, a short length of steel wire is passed through the end hole and the ends are wrapped around the antenna wire. Another short section of steel wire joins the four arms at the center hole.

This antenna should be asssembled in a clear, flat area such as a driveway. A firm support at each end of the assembly area is required. Four fiberglass spreader assemblies are required; the two center spreader assemblies are connected with insulating ropes to take the strain off the resistor-inductor assembly. It's a good idea to start at one end of the antenna and work towards the other, with the first wire (preferably the center one) strung tightly between the supports. The remaining wires can be built up around the taut center wire.

When the assemblies are completed, short lengths of copper bonding wire connect the outer cage wires to the central wire at each spreader assembly. If the wires are all pre-cut, the spreaders can be adjusted after final assembly to make up for minor inaccuracies in construction.

the fixed network

The network coil is composed of 50 turns of No. 18 wire closewound in the center of a 14-inch section of one-inch (outer diameter) PVC plastic tubing. The ends of the coil are secured by passing them through holes drilled in the form. Six 2.2K, 1-watt resistors connected in parallel (366 ohms, 6 watts) are placed across the coil. The coil assembly is mounted in the center of the two spreader assemblies and secured in position with nylon cord. The wires of the cage sections are connected to the coil ends. The whole assembly should be protected from the weather.

---

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the matching transformer

The broadband matching transformer is shown in **fig. 2**. It has a bifilar winding (2 wires). Note that one of the windings is tapped. Take two 4-1/2 foot lengths of two different color No. 18 flexible hookup wires and twist them together by holding the wire ends in a vise and placing the other two ends in a chuck of a hand drill. Keeping the wires taut, wind the drill to twist the wires until there are about two turns per inch. Wind 24 turns of the twisted pair on an Amidon T-200-2 powdered-iron core.** Keep the turns pressed against the surface of the core. Connect the start of one winding to the end of the other as shown in the drawing.

Mark one end of the dual winding with a spot of paint and consider this end the ground end. Count around the core 18 turns from the ground end, then carefully cut one wire of the winding and scrape the exposed ends. Tin the ends and solder them together with a *short* length of wire, making a loop in wire for a tap point. The transformer is now finished and can be mounted, with coaxial fittings, in a weatherproof box at the base of the antenna. The ground end of the transformer is attached to the ground rod and appropriate ground radials or water pipes.

installing the antenna

Because this is an experimental antenna, it would be nice to have the input end at or near ground level for a convenient adjustment and the free end up in the air. To achieve this, the sloper configuration is suggested. The original antenna was run from near gound level at a 45-degree angle to a point 30 feet high. Tests with the antenna in this position showed that it had a nearly-omnidirectional pattern across the operating range and that the antenna was capable of operating at, or slightly above, the 100-watt power output level of the test transmitter, a Drake TR-7. Measurements indicated the antenna feedpoint

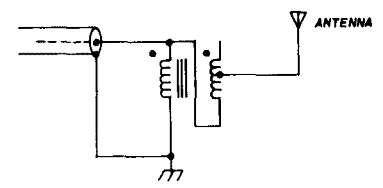

fig. 2. The matching transformer is wound on an Amidon T-200-T powdered-iron core. Twenty-four bifilar turns are used. One winding is tapped for the antenna connection 18 turns from ground. Note that the tap is on the winding not connected to the coaxial line. (See text for details.)

was capacitive below 5 MHz, rising to an inductive peak at 8 MHz, near zero-reactance between 9 and 13 MHz, and that it presented a gradual rise in inductive reactance above 13 MHz. The mean input impedance over the operating range was about 154 ohms.

On-the-air tests were run on all bands (including 30 meters) with good results. Only daylight tests were run on 160 meters, with excellent signals reported out to 50 miles or so. Best of all, no antenna tuning unit was required, making band changing for those with solid-state equipment a breeze.

a vertical version of the VK6 broadband antenna?

This is an interesting antenna for the ham-experimenter who has the time for and interest in new concepts. How might the original design, a wire cage, be modified to form a self-supporting vertical broadband antenna? Such a device would be a popular addition to any ham station, particularly where it would be difficult to erect any other form of efficient broadband antenna. I'll be anxious to hear from anyone who works with this one. (Thanks to VK6IM and VK6YX for refining the concept of the Australian dipole!)

the 4Z4RS two-band quad loop

More and more Amateurs are discovering that the single quad loop makes an excellent and inexpensive single-band antenna. It has the radiation pattern of a dipole and provides a small amount of gain over a dipole mounted at the mid-point height of the loop.

Dave, 4Z4RS, has modified his quad loop to work on two bands without the need for an antenna tuner or balanced feed (**fig. 3**). His design is for 20 and 15 meters. The loop perimeter is one wavelength, 70 feet 9 inches, (21 meters) on 20 meters. The top of the loop is open and a shorted one-half wave, 20 meter open-wire stub is attached at this point. The length of the stub is 34 feet 8 inches (10 meters).

The loop is fed at the bottom by two gamma matches which are parallel-connected to a coaxial transmission line. The whole assembly, including the gammas, is made of wire and the loop is maintained in position by nylon ropes attached to the side points.

General dimensions for the gammas are given in the drawing. The length of the gamma wires and the value of the capacitors determine the degree of match. For adjustment, the 20-meter gamma capacitor is set at minimum capacitance and adjustments are made to the 15-meter gamma to reduce the SWR on the transmission line below 1.5-to-1. The connection between the gamma and the loop is made by a copper clip which can be moved about to

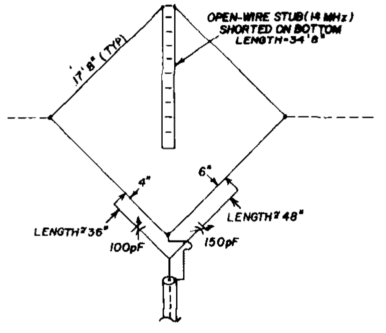

fig. 3. The 4Z4RS two-band quad loop antenna for 20 and 15 meters. Loop is cut for 20 meters and has shorted stub which acts as electrical switch for 15 meters. Parallel-connected gamma matches provide good match to coaxial line. Approximate gamma dimensions are shown.

**Red coded, powdered-iron core, permeability = 10. Two inches O.D., 1.25 inch I.D. A_L value = 120. Available from Amidon Association, 12033 Otsego Street, North Hollywood, California 91607.

an easier approach to mastering the Morse keyboard

Learn to type and CW is yours

When the semi-automatic paddle key (better known as "the bug") came along it allowed for code speeds far higher than had ever been possible with a hand key. Its arrival heralded a new era in which CW communications could be exchanged in a shorter period of time and CW was widely accepted among hams. Then came the electronic keyer which, because of its ease of operation and better code generation, rendered the bug almost obsolete. It, too, allowed for faster communications and remains in wide use today.

Over a dozen years ago the integrated-circuit Morse keyboard appeared in the ham journals. The next step in the progression of improved keying devices, it could not only outrun the electronic keyer in speed, but generated flawless CW with a fraction of the effort. I built one from an old typewriter and was so impressed with its operation that it seemed certain KB operating would sweep like wildfire through the ranks of CW operators, but it didn't. Few of the signals heard on today's bands emanate from KBs, despite the fact they can be homebrewed for substantially less than the cost of a paddle keyer, or even purchased ready-made. You would have to have a virtually empty junk box to spend more than $10.00 to create the required logic circuitry, and if you don't have an antique typewriter to mutilate, boards are available in the $20.00 range.

So what went wrong? Why do so few of us use these wonderful devices while the majority continue to struggle along, pushing, pulling, squeezing, or just generally batting around their paddle? The biggest drawback to the use of KBs seems to be that many hams never learned to operate a typewriter and fear that years of practicing monotonous drills would be required to achieve the same speed they have already attained on their paddles. The intent of this article is to dispel this fear.

Another reason for the lack of KB acceptance might be that unless one has the opportunity to operate one on the air, there is no way to appreciate how a KB can add to your enjoyment of CW. Most operators have a desire to increase their code speed, but when listening to the speed merchants who QRQ at 80 or better WPM, believe that such a feat is beyond them and must require an inborn gift or talent. This just isn't so . . . you can do it *in less time*, and with less effort than you may think.

keyboard operation

Few articles have appeared on the use of a basic KB, so a short refresher might be in order. When a key is depressed it closes a circuit, one way or another, instantly causing the logic circuitry to begin generating the particular character associated with that key *plus* a space. Once started, the logic continues to generate the character even though the key switch is opened before the character is completed. If the key is held down continuously, the character will repeat continually, with perfect character generation and flawless spacing between characters. The logic can produce only one character, plus its space, at a time. The key switches are locked out, or not enabled (if you prefer IC jargon), until that specific character is completed.

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Therefore, for smooth CW operation it is only necessary to depress the next succeeding character before the character being transmitted is completed. The longer spacing required between words is merely a pause in typing when no key is depressed.

The preceeding paragraph relates to basic KB logic without so-called memory capability. In recent years some of the logic circuits have been expanded to allow you to type continuously into memories or buffers at a somewhat erratic rate, and have the output appear a few seconds later, automatically smoothed of any timing inconsistencies that occurred in the input.

There is little benefit in being able to type Morse at a high rate of speed if your receiving ability is not equal to it. If there was ever a secret weapon for increasing your receiving speed, it's the basic KB. It may sound like something out of the Twilight Zone, but consider this: young children without the slightest knowledge that Morse characters are formed from dits and dahs have become proficient at Morse by listening to their own typing on a KB and recognizing the various characters by sound only. Most of us learned code the hard way. First dots, dashes, and timing had to be understood and then the characters were memorized. Finally the dots and dashes made letters into words. In transmitting Morse by means of a key or paddle, the brain is called upon to recall all that it learned about the formation of characters. The KB, which makes a sound for the character the instant it is keyed, eliminates that whole process.

In essence, the KB is a tireless teacher that communicates directly with your subconscious with no effort on your part. It follows that, as your typing speed increases, so does your ability to receive, because the brain is no longer required to busy itself with character formation but instead can concentrate on hearing what has been transmitted. You have only to think the letter A, strike that key, and immediately hear dit-dah. The faster you type, the faster you go.

In the early days of keyboarding it was discovered that contrary to popular notion, the speed at which CW could be received was not so much limited by the ability of the ear and brain to comprehend, as much as it was the limitation of the sending device to shape legible code. Few operators have the dexterity or endurance to operate a paddle at sustained high speeds and maintain consistency.

For the most part, ham CW is recreational communication. We're not operating aboard the Titanic, grinding out vital messages with a letter-perfect copy for the captain. In fact, we don't have to make any copy at all; just a few notes will keep a QSO going. With the KB as a teacher, not only does the brain recognize the sound of individual letters at increasingly higher speeds, but a point is reached where you suddenly realize that instead of hearing the individual letters that make up a word, you hear the complete word itself. Morse at that time becomes similar to another language, as if you were hearing French, Spanish, or some other language in which you are fluent. You might be hard-pressed to make a running verbatim translation, but the complete thoughts come through loud and clear.

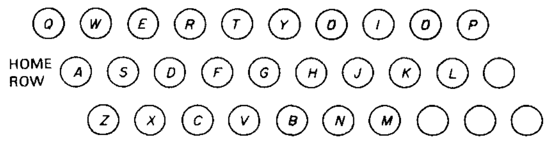


fig. 1. Letters as they appear on the standard keyboard (Qwerty).

Thus, the KB in and of itself can create a whole new world of improved CW communications for anyone who is interested. With a basic KB and the ability to use it, increased Morse proficiency is assured.

keyboard arrangements

Fig. 1. shows the letter characters as they appear on a standard KB. Numerals, punctuation marks, etc., have been omitted. This alphabetical monstrosity is a hodge-podge of letters that could have been set up by any moderately intelligent chimpanzee. Its only redeeming feature is that the first six letters on the top row can at least be pronounced and hence give rise to its name: the Qwerty keyboard. One might like to think that there was some profound scientific basis for the arrangement of the letters, but none has ever been uncovered. Instead, it apparently evolved in 1872 when some inventors were working the bugs out of a typesetting machine. It seems that the type slugs were stored in a basket and suspended on wires to permit movement. These wires often became entangled, particularly when the most frequently used keys were struck. To solve the problem, KB designers simply separated the most frequently used letters from each other, placing less frequently used keys between and among the more useful ones.

Take a look at **fig. 2**; at first glance it may look like the work of another chimp. This KB arrangement, based on the use of letters in the English language, was developed in 1932 by August Dvorak, a professor of education at the University of Washington at Seattle, with funds provided by the Carnegie Corporation (who knew there just had to be a better way to punch a KB). It is generally known as the Dvorak Simplified Keyboard, or DSK. Dvorak spent the grant money well

by studying the most-used English words, and how best to type them with the strongest fingers, while minimizing finger travel. His KB layout was proven to be superior to Qwerty in every respect, and his students ran circles around Qwerty experts in learning time, speed, and accuracy. Unfortunately, the DSK failed to catch on because of the inertia that arises when one must make a change from the norm. (Will present-day textbooks ever admit that electrical current doesn't flow from positive to negative?) Much progress is stifled because of the human tendency to resist change.

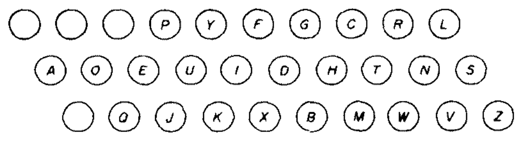


fig. 2. Letters as they appear on the Dvorak simplified keyboard (DSK).

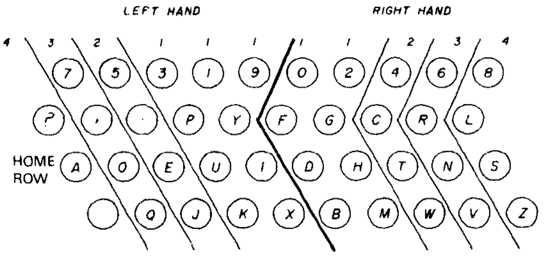


fig. 3. Recommended fingering of the DSK.

To touch-type on any KB, the operator positions his or her hands with the fingers over the center, or "home", row. Any departure from the home row requires effort and concentration. Dvorak placed all the vowels and the most used letters on the home row. Other letters are within reach of the strongest fingers in accordance with the frequency of their use. Think up some common words and mentally type them on the DSK. Try the same words again on the Qwerty board, and make your own decision as to which is easier. You can type about 3000 common words without ever leaving the DSK home row. The DSK system allows 70 percent of the typing to be done on the home row, 22 percent on the upper row, and 8 percent below.

Fig. 3 shows the recommended fingers to use with each key, as well as where Dvorak located the numerals and some punctuation marks. The stronger fingers do double-duty at the center of the board, where the most-used letters are located. The weaker pinky-fingers loaf along and don't often leave the home row. For hamming purposes, location of the numerals is somewhat arbitrary. Most boards have many more keys than shown, so additional punctuation marks and CW procedural signs can be placed on unused keys wherever they seem most convenient.

The DSK is not completely dead, and many typewriter manufacturers will supply them at additional cost. Self-study books are available, but not necessary. The board can be learned quite readily by placing the fingers over the home row, with pinky fingers over the A and S. Stroke the keys and return the fingers to their home position each time they are required to leave it. As each character is learned, cover its key-top with a piece of tape to avoid cheating by peeking.

A few cautions: don't practice by typing from printed material. You can make good progress this way but somehow the printed words flow directly from the eyes to the fingers without stopping in the brain. The brain must be involved, or when QSOing your typing mind is quite likely to go blank. Second, when practicing always type a little faster than you think you can. If you slow down, the choice factor rears its ugly head and errors are much more likely to occur. Be sure to use a keying transistor in the output of your logic circuit, as mechanical relays or magnetic reeds will not move fast enough when you get into high gear. And by all means use a break-in system that allows you to hear incoming signals between your dits.

There is a rhythmic difference between typing alphabetic letters and typing Morse characters, which probably prompted development of the memory learning systems previously mentioned. In alphabetic typing the keys are all stroked at the same speed, in an even cadence. With CW, however, you can linger a while on the longer characters, such as Y, before going on to the next character. The shorter characters, such as E, must be vacated as soon as they are struck and the succeeding character immediately activated. This demands irregular finger movements but, by monitoring the output, perfect timing is readily achieved on the DSK. Storage systems are an unnecessary adjunct to the basic KB, and defeat its usefulness in gaining higher receiving speeds.

Most people are able to learn the locations of all the letters in about two weeks. That's the hard part, but once accomplished you're into a typing system that will have you typing more than 100 WPM in less than a year. And believe it or not, there are ears out there that will be able to copy you.

ham radio

VHF/UHF WORLD

Joe Reisert W1JR

VHF/UHF exciters

It wasn't too many years ago that VHF/UHF'ers used AM, FM, and CW with tube or varactor multiplier transmitters almost exclusively. While signals weren't always that stable, neither were the receivers; VHF was considered a quiet refuge away from the QRM on HF and was considered good only for local ragchews or experimentation.

With worldwide DX becoming commonplace on the frequencies above 50 MHz,[1] those days are now a distant memory. Most modern receivers have low noise figures and are very frequency stable.[2] Hence most weak-signal VHF/UHF'ers now prefer to use either CW or SSB at the flip of a switch. DX is suddenly possible where it was never expected to appear, and as a result, the old tube or varactor multipliers with their chirps and key clicks are now being replaced.

With this in mind I thought it would be a good time to review some of the techniques presently being used as exciters on the VHF/UHF bands and to show some simple circuits you can build in your shack. Hopefully this will convince you that trying out some of the higher frequency bands doesn't have to be expensive or complicated. Who knows? Maybe you'll find that there's still some room for exploration and experimentation, and that you, too, can contribute to the state-of-the-art. New propagation modes are rarely discovered without activity!

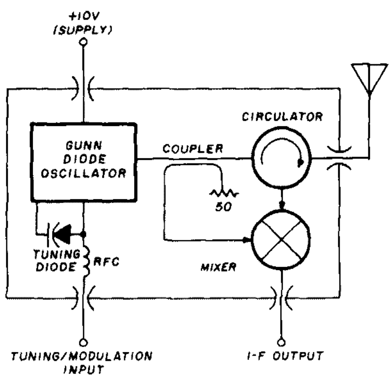


fig. 1. Typical Gunnplexer™ configuration.

exciter techniques

There are basically three ways to build a VHF/UHF exciter: first, a free-running oscillator; second, a multiplier; and third, a transverter. Although the free-running oscillator may seem ludicrous, don't forget that this technique, shown in **fig. 1**, is still very popular on the 3 cm (10 GHz) band. Gunn-Plexers™[3], transceivers that use a single common Gunn diode oscillator as both a varactor-modulated FM transmitter and as the local oscillator for the receiver downconverter, are in widespread use. However, this scheme is not stable enough for narrowband work unless it is frequency-locked to a reference.

Tube multipliers (**fig. 2**) used to be very popular. Since the oscillator was usually at some low frequency (for ex-

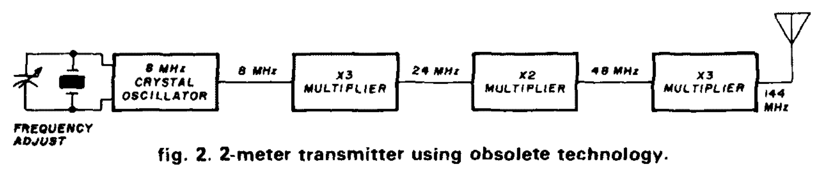


fig. 2. 2-meter transmitter using obsolete technology.

ample, 8 MHz), the only way to adequately control frequency was by using a crystal oscillator. Some VHF/UHF'ers added a variable capacitor across the crystal to slightly "rubber" the frequency. However, especially when using *vacuum tube* local oscillators, the frequency stability was poor. Hence, in the "good old days," some EME'ers even buried their oscillators in the ground in order to make their stability more acceptable!

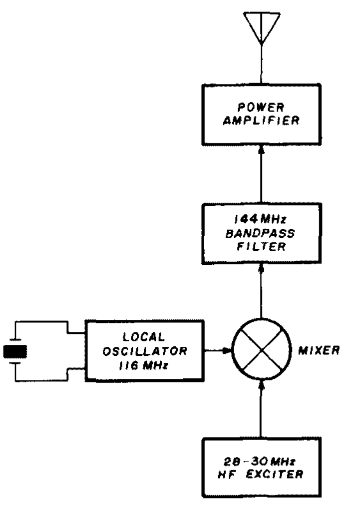


fig. 3. Typical 2-meter transverter.

Most of these problems disappeared when transverters (**fig. 3**) became popular, especially when transistor oscillators were used. A transverter is basically an upconverter or mixer working inverse to the normal receiver downconverter. The primary advantage is that the local oscillator can run continuously and thereby attain a high order of stability. A low frequency transmitter (driver) is mixed with this LO and is usually the main frequency-determining element. Since HF rigs are now quite stable up through 30 MHz, they make a natural driver for such an application.

transverter basics

There are basically two types of transverters: low level and high level. In the low level type, the signal from the driver is typically less than one watt; 1 to 10 milliwatts is most common. This approach requires considerable gain, but is easy to obtain using inexpensive solid-state devices. The high level type has most often been used with *vacuum tube* mixers such as the 5894, 6360, 6939, or 2C39/7289.

Although I started out on VHF/UHF using 6360s and 6939s, I soon tired of all the drifting and warm-up problems associated with vacuum tubes. Furthermore, I didn't appreciate all the different high voltages and the heat associated with tubes. So I spent many hours developing solid-state approaches, especially those that were low in cost and easy to duplicate. What eventually evolved was a compatible system for all the VHF/UHF bands through 13 cm (2304 MHz), using only 1 milliwatt of drive from my HF exciter and providing 200 to 500 milliwatts of output power on each of the bands mentioned. Because space is necessarily limited, I'll discuss only the 2 meter through 70 cm (432 MHz) circuits in this column. A block diagram of their scheme is shown in **fig. 4**. (Is there any interest in circuits for any of the other bands mentioned? *Let me know.*)

circuit details

The heart of any transverter scheme is the mixer. Single diodes, bipolar transistor and dual gate FET mixers were all discarded for various reasons in the early stages of development and

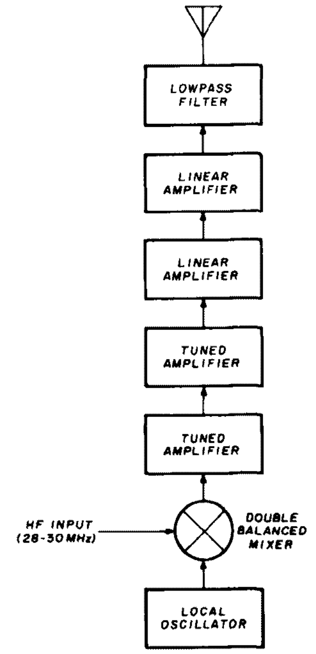


fig. 4. Block diagram of the 144/220/432 MHz transverter discussed in text.

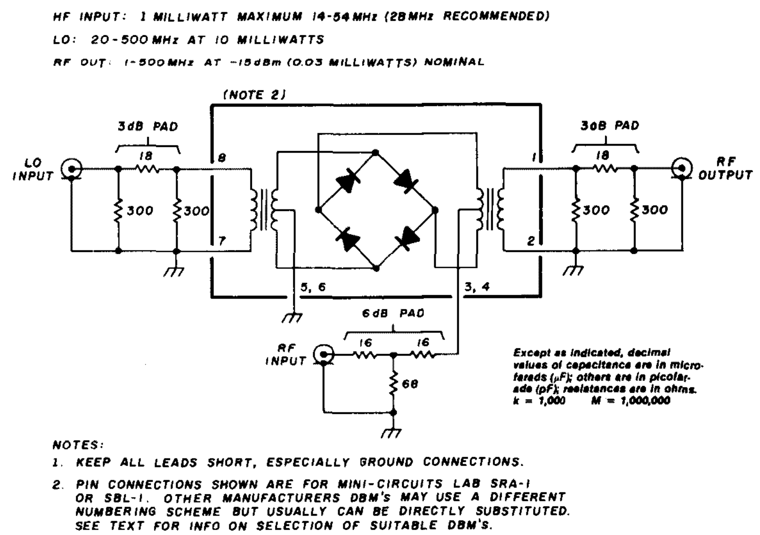


fig. 5. DBM up-converter for VHF/UHF transmitters.

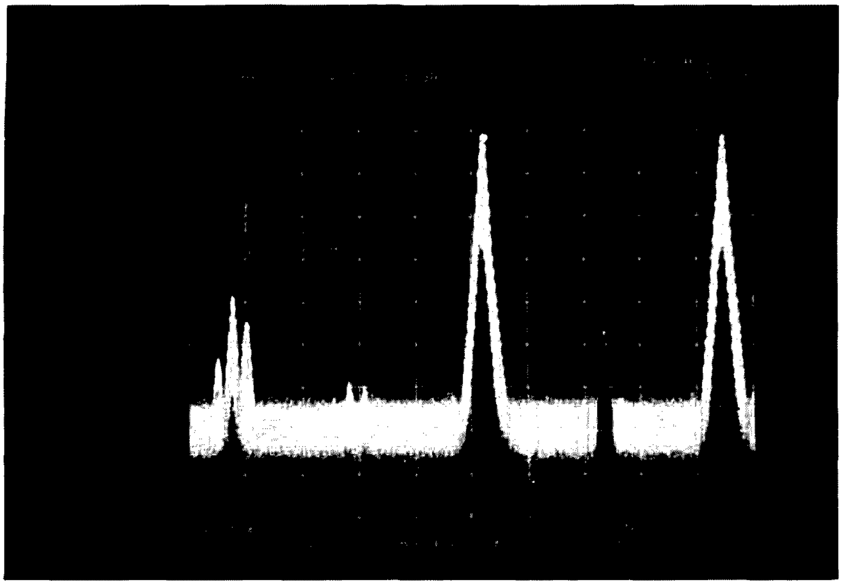


fig. 6A. 2-meter transmitter converter output spectrum from DBM.

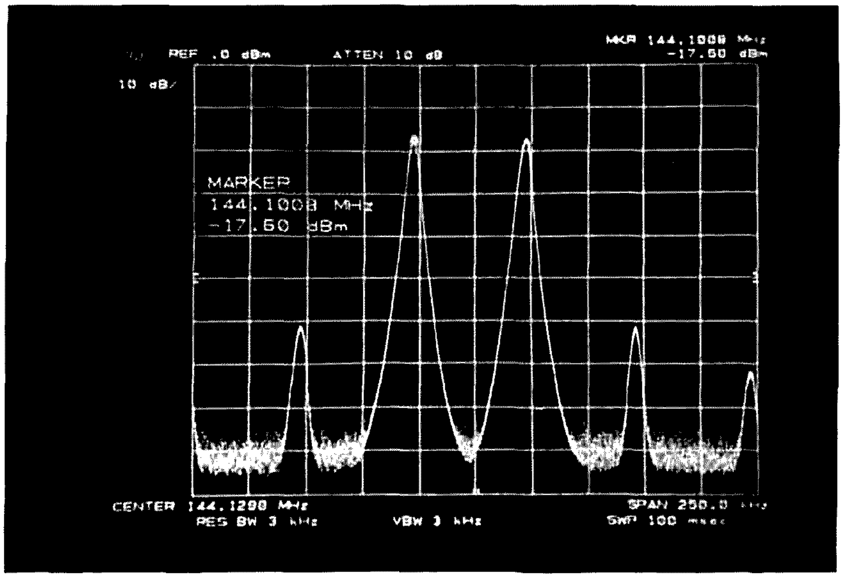


fig. 6B. 2-meter transmitter converter output spectrum at desired frequency range.

eventually replaced with the ubiquitous DBM (double balanced mixer), which is 50 ohms at all ports. The common low level (5 milliwatt LO) DBM was chosen because higher level DBMs cost more than twice as much as low level DBMs and require 10 dB more local oscillator power. The circuitry is quite similar in concept to the receiver downconverter described in last month's column.[2] The main differences are that the receive input is now the transmitter output, and the IF output is now the input (see **fig. 5**).

Early in the design phase I decided that the IMD (intermodulation distortion) in the mixer had to be quite good — (much greater than 30 dB down) — because each successive amplifier stage would degrade the final performance by a small amount. The use of 1.0 milliwatt drive from the exciter and

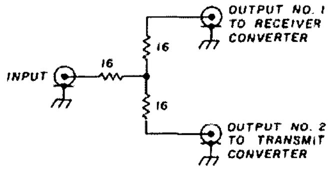
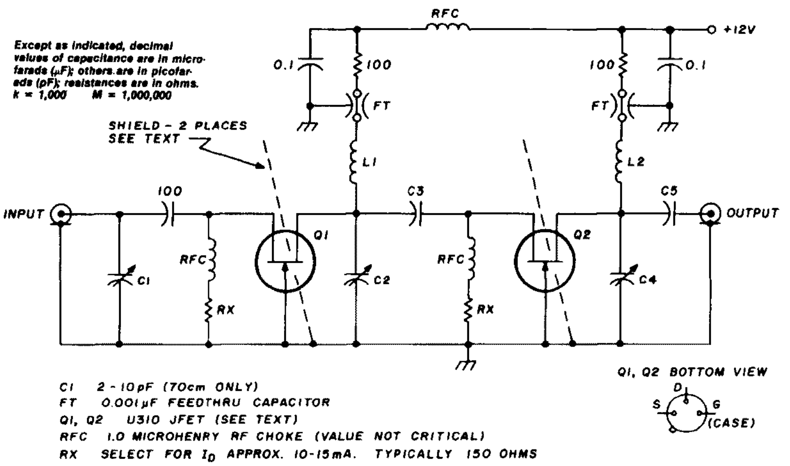


fig. 7. Simple two-way 50 ohm power splitter.

a 6 dB attenuator pad on the IF port provided the optimum balance between the IMD desired and the conversion loss in the mixer. A typical broadband two-tone output spectrum from the mixer of 2-meter transverter using this scheme (see **fig. 5**) is shown in **fig. 6A**. A close-in output spectrum is shown in **fig. 6B**. Note that the output level is approximately 0.02 milliwatts and that the IMD is down over 40 dB, a good start for a low-level transverter. Again, the Mini-circuits Labs SRA-1 was much better than their SBL-1. Additional details on the DBM selection are described in reference 2.

The LO (local oscillator) is provided by a solid-state overtone oscillator and multipliers where required, with typically 10 milliwatts output similar to the one described in last month's column.[2] If transverter operation is desired, a simple minimum-loss "T" pad (**fig. 7**) can be substituted in place of the output attenuator in the LO and thus provides the required dual outputs. The LO will not be described here since it was well documented in last month's column.

The output of the DBM is the desired signal plus its image and feedthrough from the LO and driver (**fig. 6A**). Therefore some filtering is required. Until recently I used a bandpass filter following the DBM, but I now find that if you use narrow-band tuned amplifiers with good input VSWR, extra filtering is usually not required. My favorite tuned amplifier uses two simple grounded gate U310 JFETs (**fig. 8**). The U310 is usually thought of as a low-noise receiving preamplifier, but it makes an excellent stable linear amplifier with a typical



	144 MHz	220 MHz	432 MHz
C2, C4	2-18pF	2-10pF	2-10pF
C3, C5	3.0pF	2.0pF	1.0pF
L1, L2	4T, NO.20 AWG 1/4"(6.4mm) ID x 1/2"(12.7mm) LONG	3T, NO. 20AWG 1/4"(6.4mm) ID x 1/2" (12.7mm) LONG	1.0"(25.4mm) NO. 20 1/4" (6.4mm) ABOVE PC BOARD

fig. 8. Low-level JFET linear amplifier schematic.

gain of 10-12 dB per stage and over 10 milliwatts linear output. Input VSWR is typically 1.5:1 without matching networks, and half-power bandwidth is typically 2 to 3 percent of the operating frequency. *Do not substitute any other JFET for the U310 since others may not have the gain and/or stability necessary for this circuit to perform to specification. The J310, for instance, is the same chip in a plastic package, but it is potentially unstable above 100 MHz.*

The output of the two-stage U310 circuit is typically 1 to 5 milliwatts if used with the DBM as shown in **fig. 5** — hardly enough to make many DX contacts! Hence more gain is required. A broadband feedback type linear amplifier using constant current source biasing (**fig. 9**) was designed. This amplifier design has moderate gain (typically 10 to 15 dB per stage) and good linearity with typically 100 to 250 milliwatts on SSB and 0.5 watts output on CW. It uses low-cost ($5.00 - $10.00) CATV type stud-mounted transistors. Most of my linear amplifier circuits use the Microwave Semiconductors Inc. 82091 but other similar devices such as the Acrian CD1899, NEC NE74020/2SC1251, Solid-State Microwave SD1005, and TWR LT2001 will also deliver similar performance. This kind of device is typically a 1-watt output transistor with a 1.5-2.5 GHz F_T that has been optimized with internal emitter ballasting resistors for good linear operation in CATV type amplifiers. Class "C" type transistors are not recommended because they are usually highly non-linear.

The constant current source[4] allows the emitter to be at a low impedance or directly grounded while providing excellent controlled collector current in the power amplifiers. The 39 and 13 ohm resistors (R3 and R4) "set" the stage current at nominally 50 and 150 mA respectively regardless of the device. A simple lowpass filter on the output stage eliminates any harmonic output if this is the final stage of your transmitter.

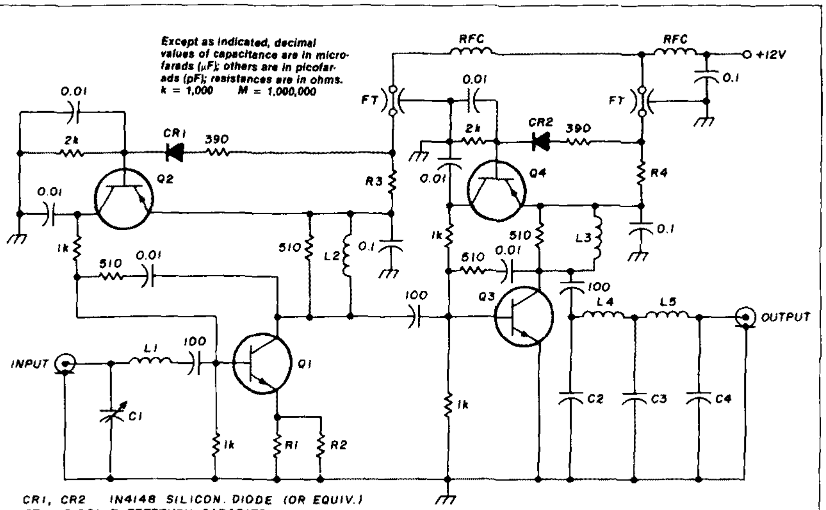
The above amplifier was specifically designed to work on 12 to 13 volts DC. Greater output power is available if the supply voltage is increased and the biasing resistor values are changed, but this is beyond the scope of this article. Also, if you are fortunate enough to have access to typical general purpose broadband hybrid amplifiers or CATV amplifier chips, these may also work; but they must be evaluated on a one-on-one basis because they come in a variety of gains and power output, and may therefore require different power supply voltages.

construction notes

The leads on the DBM — particularly the grounds — should be kept as short as possible. If you substitute another DBM, check the pin connections carefully since some manufacturers number the pins differently. I prefer to mount the DBM in its own shielded box with 3 "BNC" type connectors. If you do this, you can use the same mixer for any band within the specified frequency range of the DBM by just changing the LO!

The low level U310 amplifier is quite easy to build. Again I suggest that this circuit be built in its own shielded box; I suggest the Bud CU-124 or any equivalent cast aluminum box. First cut a piece of double-clad PC board to fit inside the cover of the box. Then drill the holes for the input/output connectors ("BNC" preferred), the transistors and feed-through connectors with the PC board in place. Next attach the feed-through capacitors and input/output connectors which hold the PC board in place. The U310s are inserted upside-down in a round hole (use a No. 11 drill) approximately 0.191 inch (4.85 mm) in diameter. The can is then quickly soldered to the PC board ground. Next the rest of the RF circuitry is soldered in place; all grounds can be easily attached with short leads to the PC board. Next, shields approximately 0.75 inch (19 mm) high and the width of the PC board are made from the same type material. They are first notched to pass the U310 leads and then sweat-soldered directly to the PC board across the JFET, providing input-to-output isolation. The 100-ohm resistors and associated circuitry in the drain lead can be mounted on top of the box on standoff insulators.

The output amplifier is constructed



CR1, CR2 1N4148 SILICON DIODE (OR EQUIV.)
FT 0.001µF FEEDTHRU CAPACITOR
L2, L3 5T NO. 20 AWG ON 1/4" (6.4mm) DIAMETER x 0.5" (12.7mm) LONG
Q1, Q3 SEE TEXT
Q2, Q4 2N2905 OR EQUIV HIGH GAIN PNP TRANSISTOR
RFC 1.0 MICROHENRY RF CHOKE. VALUE NOT CRITICAL
R1, R2 10 OHMS, 1/4 WATT
R3 39 OHM, 1/4 WATT
R4 13 OHM (THREE 39 OHM, 1/4 WATT, RESISTORS IN PARALLEL)

	144 MHz	220 MHz	432 MHz
C1	NOTE 1	NOTE 1	1-10pF
C2, C4	24pF	18pF	NOTE 1
C3	47pF	36pF	NOTE 1
L1	NOTE 1	NOTE 1	1/2" (12.7mm) LEAD ON THE 100pF CAPACITOR
L4, L5	6T, NO. 24AWG ON 0.1" (2.5mm) DIAMETER 0.25" (6mm) LONG	3T, NO. 24 AWG ON 0.1" (2.5mm) DIAMETER 0.125" (3mm) LONG	NOTE 1

fig. 9. Medium-power linear amplifier.

in the same type box and in the same manner as the U310 stages. In this case the linear transistors are stud-mounted. First drill the hole to clear the transistor stud. Then enlarge this hole only on the PC board to approximately 0.32 inch (8 mm) in diameter to allow the body of the transistor to pass through. Next bolt the connectors and transistors to the cover. Again, all wiring is point-to-point and grounds are kept to a minimum by soldering directly to the PC board. The box cover acts as a heatsink for the power transistors.

testing

The DBM doesn't really require any testing. First connect the LO with 10 milliwatts to the proper input. Then set your driver output for 1 milliwatt. With the modern solid-state rigs an auxiliary output with typically 10 to 500 milliwatts is usually available. To insure that the driver is operating properly, it is best to operate at normal levels so that the ALC is activated. Therefore I recommend that you place an external "T" or "Pi" pad of the required value on the driver output to guarantee that only 1 milliwatt is present at the mixer input connector rather than just turning the gain control down. If you do not have a low-power output connection, use a suitable dummy load and capacitively couple out the proper power level.

Now connect the RF output from the mixer through a short coax cable to the two-stage JFET amplifier. Apply the proper power and input frequency from the exciter and peak each stage for maximum output. There may be some interaction between the stages, so repeat the peaking until you can't obtain any more output power. If you have a power meter, measure the output level. It should be between 3 and 10 milliwatts, less at the higher frequencies.

Next connect the low power stage through a cable to the medium power stage and put a power meter on the output. Lower the exciter drive so that the output is about 100 milliwatts and peak the input trimmers (if used) for maximum power. Also repeak the last U310 output stage. Apply the proper drive and measure the output power. If it is greater than 500 milliwatts, you may want to reduce the exciter input slightly, or better yet, place a low value pad between the JFET and bipolar stage to preserve linearity. This output level should be sufficient to work locals. If more power is desired, there are many suitable circuits or commercial amplifiers available to get you to the desired power level.

summary

In this month's column we have discussed several approaches to building VHF/UHF exciters, but have recommended the transverter for its stability, linearity, and versatility. Several circuits and construction techniques have been presented to allow you to "roll your own." You no longer have a good excuse for not trying these frequencies. Furthermore, if you use this method, you can build on more gain and power later without having to start all over again!

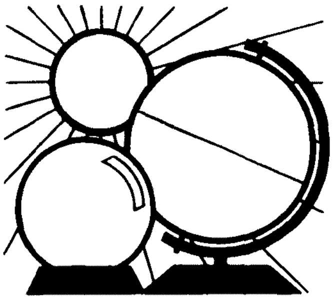
references

1. Joe Reisert, W1JR, "VHF/UHF World: the VHF/UHF Challenge," *ham radio*, January, 1984, pages 42-43.
2. Joe Reisert, W1JR, "VHF/UHF World: VHF/UHF Receivers," *ham radio*, March, 1984, page 42.
3. James Fisk, W1HR, "Solid-State Microwave RF Generators," *ham radio*, April, 1977, pages 10-22.
4. Bill Smith, KØCER, "The World Above 50 MC: The W6FZJ Wide-Band Low-Noise Preamplifier," *QST*, November, 1972, page 112.

VHF/UHF coming events

April 7-8: Best EME Weekend
April 16: ARRL 2-meter Sprint Contest
April 21: Predicted Peak, Lyrids Meteor Shower (0800 UTC)
April 24: ARRL 220 MHz Sprint Contest
April 27-29: UHF/VHF Conference, Dayton Hamvention (Contact WA8ONQ for information.)

ham radio



DX FORECASTER

Garth Stonehocker, K0RYW

last-minute forecast

April is an equinoctal (i.e., the length of day and night is approximately equal) month. Consequently, it's reasonable to expect geomagnetic-ionospheric disturbances, with enhanced transequatorial propagation on the 10 to 30-meter bands. Disturbances are most likely to occur between April 1-5, 8-13, 16-19, and 25-28.

DX openings for the higher frequency bands, 10-30 meters, should be best about the 8th or 21st; because solar flux and activity are becoming more difficult to predict for this part of the forecast, be sure to monitor WWV at 18 minutes after the hour for the most current solar flux information.

The lower frequency bands are expected to be very good this month. But this will be one of the last months before the general northern hemisphere thunderstorm noise buildup becomes noticeable.

The perigee of the moon's orbit (for moonbounce DX) is on the 13th; the moon will be at full phase on the 15th. There will be a short meteor shower, the Lyrid, on April 20-22 with a rate of five per hour — hardly much help for meteor-scatter DX. But a bigger shower, the Aquarid, starts before the end of April, peaks on May 5, and ends in mid-May. Its rate is 10 to 30 per hour.

solar cycle effects

In September's column I reviewed the solar cycle and presented a formula relating critical frequency, foF2, in the mid-latitudes to solar indices. Its significance is that the foF2 is the most variable factor of radio signal propagation. The maximum usable frequency (MUF) for the path is approximately 2.5 times the critical frequency.

How has the solar cycle/MUF cycle decline progressed so far? From early 1982 until December, 1983, the sunspot number (SSN) declined quite rapidly — about 40 units per year. Since then a more leisurely decrease (about 20 units per year) has been reported. This rate is expected to continue until the SSN "bottoms out" in 1986.

The SSN for May, 1983, was down to 77.1. The energy output of the sun (solar flux) decreased along with SSN, and with that the foF2 and MUF have been dropping, too.

You probably noticed the drop, particularly by November and December of last year; low radio flux and solar activity for nearly two weeks in a row made MUF decreases very noticeable. Ten meters was not open very long — if at all from some locations — by the end of the period. The 15-meter band wasn't too much better. Even the 20-meter band closed earlier in the afternoon. Several of the beacons I monitor dropped out early in the day, and my WWV monitoring and data-gathering activity had to drop down a standard frequency band for that two-week period.

Another solar cycle effect noticeable over the last few winter months was longer distances and better signal strengths on the lower frequency bands during the daytime; this was particularly noticeable on 40 meters. The lower level of D region ionization, near sunspot minimum, combined with the sun's being in the southern hemisphere, caused lower signal attenuation. Lower ion densities mean less signal absorbed in the D region. The effect should be somewhat effective even this summer, and next winter, even more so, as the SSN will have decreased a bit more. However, even the lowering of the attenuation or lowest usable frequency (LUF) does not make up for the decrease in maximum usable frequency. Because the extent of the frequency range, LUF to MUF, decreases during these SSN minimum years, we should try to use the 160-meter band more effectively to compensate for loss of propagation on the higher frequency bands.

band-by-band summary

Ten meters will be open to the south and southeast for a short period before local noon, to the south at noon and to the southwest in the afternoon. The openings will be longer when the solar flux is at its 27-day cycle maximum. Even better transequatorial one-long-hop conditions will occur during disturbed periods. Tune in WWV at 18 minutes after the hour and note particularly the geomagnetic field status announcement.

Fifteen and twenty meters, almost always open to some part of the world, will be the main daytime DX bands. Twenty should stay open on long southern paths into the night, while 15 will drop out in the late afternoon. Operate 15 first and move down to 20 meters. DX is 5000 to 7000 miles (8000 to 11,200 km) on these bands and one-long-hop transequatorial propagation is also possible as on 10 meters.

WESTERN USA

GMT	PST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	4:00	20	15	15	10	15	10	10	15
0100	5:00	20	15	15	10	15	10	10	15
0200	6:00	20	15	15	10	15	10	10	15
0300	7:00	20	20	20	10	20	10	10	15
0400	8:00	20*	20	20	10	20	10	10	15
0500	9:00	15	20	20	10	20	10	10	15
0600	10:00	15	20	20	10	20	10	15	20
0700	11:00	15	*20*	20	15	20	10	15	20
0800	12:00	15	*30*	20	15	20	15	15	20
0900	1:00	*30*	*30*	*20*	15	*30*	15	15	20
1000	2:00	*30*	*30*	*20*	*20**	*30*	15	20	20
1100	3:00	*30*	*30*	*20*	*20*	*30*	*20*	20	*30*
1200	4:00	*30*	*20*	15	*20*	*30*	*20*	*20*	*30*
1300	5:00	*30*	*20*	15	*20*	*30*	*20*	*20*	*30*
1400	6:00	*30*	20	15	20	*20*	*20*	*20*	*30*
1500	7:00	30	20	15	20	*20*	*20*	*20*	*30*
1600	8:00	30	20	10	20	*20*	20	20	*30*
1700	9:00	30	20	10	20	*20*	15	20	20
1800	10:00	30	20	10	15	*20*	15	20	20
1900	11:00	30	20	10	15	15	15	15	20
2000	12:00	20	15	10	15	15	10	15	20
2100	1:00	20	15	10	15	15	10	15	20
2200	2:00	20	15	10	10	15	10	15	20
2300	3:00	20	15	15	10	15	10	10	20

APRIL

MID USA

MST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN	CST
5:00	20	15	15	15	15	10	10	15	6:00
6:00	20	20	15	15	15	10	10	15	7:00
7:00	20	20	15	15	15	10	10	15	8:00
8:00	30	20	20	20	20	10	10	15	9:00
9:00	30	20	20	20	20	10	10	15	10:00
10:00	*30*	20	20	20	20	10	15	15	11:00
11:00	*30*	20	20	20	20	10	15	20	12:00
12:00	*30*	*30*	20	*20*	20	15	15	20	1:00
1:00	*20*	*30*	*20*	*20*	*30*	15	15	20	2:00
2:00	*20*	*30*	*20*	*20*	*30*	15	20	20	3:00
3:00	*20*	*30*	*20*	*20*	*30*	*20*	20	*30*	4:00
4:00	*20*	*30*	*20*	*15*	*30*	*20*	*20*	*30*	5:00
5:00	*20*	*30*	15	15	*30*	*20*	*20*	*30*	6:00
6:00	20	20	15	15	*30*	*20*	*20*	*30*	7:00
7:00	20	20	15	15	*20*	*20*	*20*	*30*	8:00
8:00	20	20	15	10	*20*	*20*	*20*	*30*	9:00
9:00	15	20	10	10	*20*	20	20	20	10:00
10:00	15	20	10	10	*20*	15	20	20	11:00
11:00	15	20	10	10	*20*	15	20	20	12:00
12:00	15	20	10	10	15	15	15	20	1:00
1:00	15	20	10	10	15	10	15	20	2:00
2:00	20	20	10	10	15	10	15	20	3:00
3:00	20	15	15	15	15	10	15	20	4:00
4:00	20	15	15	15	15	10	10	20	5:00

EASTERN USA

EST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE CARIBBEAN S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
7:00	20	20	15	15*	15	10	10	15
8:00	20	20	15	15	15	10	10	15
9:00	30	20	15	15	15	10	10	15
10:00	30	20	20	15	20	10	10	15
11:00	30	20	20	20	20	10	15	15
12:00	30	20	20	20	20	10	15	15
1:00	*30*	*30*	20	20	20	15	15	20
2:00	*30*	*30*	20	20	20	15	20	20
3:00	*20*	*30*	*20*	*20*	*30*	15	20	20
4:00	*20*	*30*	*20*	*20*	*30*	*20*	20	*20*
5:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	*30*
6:00	*20*	*30*	*20**	*20*	*30*	*20*	*20*	*30*
7:00	*20*	20	15	*15*	*30*	*20*	*20*	*30*
8:00	*20*	20	15	15	*30*	*20*	*20*	*30*
9:00	20	20	15	10	*20*	*20*	*20*	*30*
10:00	20	20	10	10	*20*	20	*20*	*30*
11:00	15	20	10	10	*20*	15	20	*30*
12:00	15	20	10	10	*20*	15	20	20
1:00	15	20	10	10	*20*	15	15	20
2:00	15	20	10	10	15	15	15	20
3:00	15	20	10	10	15	10	15	20
4:00	20	20	10	10	15	10	15	20
5:00	20	20	10	15	15	10	10	20
6:00	20	20	15	15	15	10	10	20

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during 'normal' hours.
*Look at next **higher** band for possible openings.

Thirty and forty meters are both day and night bands. Intermediate distances (1000-1500 miles, 1500-2200 km) in any direction is daytime DX. Nighttime DX on these bands may be expected to offer a greater distance than on 80 meters and, like 80, follow the darkness path across the sky. Reduced midday signal strengths and distances may occur on days of high solar-flux values, with 30-meter openings disappearing in the pre-dawn hours on the morning after the high radio-flux value.

Eighty and one-sixty meters will exhibit short skip conditions during daylight hours and lengthen for DX at dusk. These bands follow the darkness, opening to the east just before your sunset, swinging more to the south near midnight, and ending up in the Pacific areas during the hour or so before dawn. The 160-meter band opens later and ends earlier.

Coastal stations and those with good low-angle radiating systems will usually have the edge for working rare DX. QRN will be as low as on some nights as experienced during the wintertime DX season.

ham radio

the HP-IL serial loop

new method transfers digital data

As noted in last month's article,[1] Hewlett Packard was the driving force behind the definition and subsequent universal acceptance of the HP-IB or Hewlett Packard-Interface Bus. After having established this standardized method of exchanging parallel digital data, the company has once again set out to establish (if they haven't already) another digital data transfer method called the HP-IL, short for *H*ewlett *P*ackard *I*nterface *L*oop.

To encourage universal acceptance of this serial, considerably less expensive, and simpler convention, the company has developed the innovative HP-IL Interface kit. This kit, the simple serial data transfer convention (soon to be a standard), and various pieces of HP-IL-compatible equipment already developed are the subject of this article.

The loop concept. Table 1 compares the important features of HP's two digital data transfer methods, noting also their important differences. Unlike the HP-IB, the HP-IL is a *uni*directional transmitter of serial digital data that transmits data to the next HP-IL device. Along with the data, sufficient energy is transmitted to energize the next HP-IL device in the loop. The waiting device rests idly until energized, making battery-powered portable operation a reality. Once the data has been transmitted through all the HP-IL devices, it returns to the transmitter/controller and is checked for errors. This checking process simply compares the data initially transmitted with the data returned after completion of the loop.

Master/slave configuration. Digital data transfer schemes use a concept of a "talker" (transmitter) and a "listener" (receiver), as discussed in the previous article. With the HP-IL, one talker is designated as the controller, or master; by necessity, all other devices within the loop are slaves — regardless of whether they are inherently listeners or talkers by design. It is possible, however, to operate without a controller. For example, a voltmeter (talker) could log readings onto a printer (listener). The use of multiple controllers with prioritized levels of interrupts is also possible.

The two HP controllers most useful with the HP-IL are the HP41 handheld calculator and the HP85A computer (**fig. 1**). Since the HP41 enjoys widespread popularity, we'll discuss only calculator controller applications in this article.

Novel logic conventions. Conventional digital data, logic highs and lows, have certain inherent and distinct voltage levels: for example, a TTL high is approximately 3.2 V and a low is near ground. However, in the NRZ (non-return to zero) convention, a change in magnetic flux is interpreted as a logic high and a lack of change, as a logic low. (This is very useful with magnetic tape recording). Therefore, with NRZ, only transitions (whether they be rising or falling edges) — not levels — constitute logic states. All non-transitions are logic zeros.

The HP-IL uses an unusual but clever convention called three-level codes. This convention results in greater noise immunity, higher reliability, and less power consumption than normal two-level codes, with their narrow pulse widths that require precise one-shots or digital timers (which are difficult to accomplish with micropower IC technology).

With three-level codes, a logic high pulse is a + 1.5 V followed by a − 1.5 V and the opposite is a logic low (see **fig. 2**). This is good because if a high or low pulse occurs during an electrical interference, it must be followed by a pulse of the opposite magnitude or it will be ignored as invalid. Transmissions are each 11 bits long and occur in frames with a special encoded 1 and 0 for synchronization at the beginning of a frame. (These are designated as 1S and 0S respectively in **fig. 2**. The "S" stands for *synchronization*.) Five of the 11 bits are for device identification; that is, $2^5 = 32$, but since a state of all zeros is invalid, a possibility of 32 minus this one state yields 31, and this is precisely how many HP-IL devices can reside within the loop.

Pulse transformers. Instead of semiconductor line drivers and receivers, transformer-isolated line drivers and receivers are used. Transformers make it easier to generate the three voltage levels required with the HP-IL code. As passive elements, they require no standby power, as previously mentioned. Transformers also have no DC component and help avoid ground loops; a 100 ohm transmission line driven with a 1.5 V level requires 10 times less power than one driven with a 5 V level. With a transformer, impedance transformation is also easy,

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table 1. Comparison of HP-IL (loop) to HP-IB (bus).

		function	
model	**speed**	**data format**	**semiconductor technology**
HP-IL	20 Kbits/sec. theoretical 5 Kbits/sec. actual	serial	CMOS
HP-IB	1 M bytes/sec.	parallel in bytes	TTL/bipolar
	transmission length limit	**power consumption**	**addressing/ease of programming**
HP-IL	100 meters with twisted shielded pair, 10 meters with zip cord wiring	low, state-of-the-art in conductor and micro-power design techniques	automatic
HP-IB	20 meters (62 feet)	moderate, typical bipolar rate	conventional
	ease of physical interfacing	**connector type**	**characteristics of intended instrument group**
HP-IL	very easy, male-to-female/ data flows out of male into female	simple 2-pin male/ female plug	low cost, relatively simple
HP-IB	relatively easy	25-pin subminiature "D"	moderate cost, more sophisticated

fig. 1. HP41 and HP85A controllers.

as we will later discover when examining the HP-IL Interface kit.

Compatibility. Shortly after the announcement of anything new in electronics, a flurry of kindred devices inevitably appears — generally with no apparent standardization in mind — resulting in a total lack of compatibility among equipment. Remember the first op-amps with all sorts of packages and pin configurations? Once standardization occurred, an 8-pin minidip had to have its pins in an accepted pattern because when manufacture of a product was begun, the production line couldn't be stopped for alteration of the PC board to accommodate a different pinout. A more recent example occurred in the struggle between the two EPROM giants, Intel and Texas Instruments, when each produced electrically identical 24-pin ICs with different pin configurations.

In response to these circumstances, HP developed the HP82166A HP-IL converter (**fig. 3**), which transforms noncompatible general-purpose I/O devices into HP-IL devices. I believe this will be the key to the HP-IL's success, since this device interfaces the HP-IL with the outside world. One would hope that other manufacturers will build this capability into measurement instruments in the future rather than require the use of more costly computers to interface with the instrument.

Those readers who own home computers may be familiar with the VIA (versatile interface adaptor) or PIA (peripheral interface adaptor), which are programmable I/O ICs with ports programmable as inputs and outputs. While based on the same underlying concept as PIAs and VIAs, the HP-IL converter is much "smarter" than these devices and can recognize HP-IL instructions and data, and change both from the HP-IL serial format to an 8-bit parallel format. This is done by control logic manipulating

a transfer buffer. A buffer is no more than a group of registers (in this case 32) that holds data. It also, through its control logic, stores operating information, implements selected operating modes, and controls the flow of and the way in which data is interpreted within the converter.

Calculator as controller. Of the multiple plug-in ports on the back of the HP-41 — a feature that has intrigued many owners — four are available for plug-in RAMs,

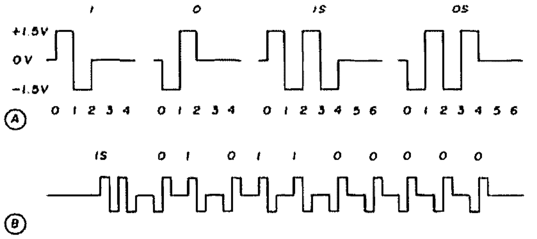

fig. 2. Three-level code (A) makes information a function of level transmissions only. An example of an encoded message — the Send Data ready frame — is shown in (B).

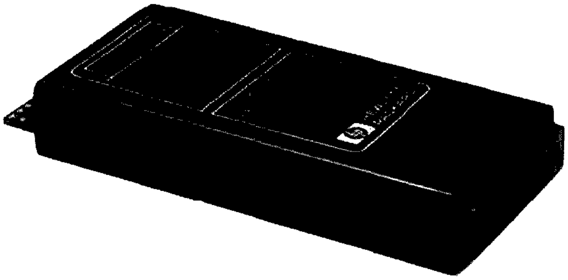

fig. 3. HP82166A HP-IL converter.

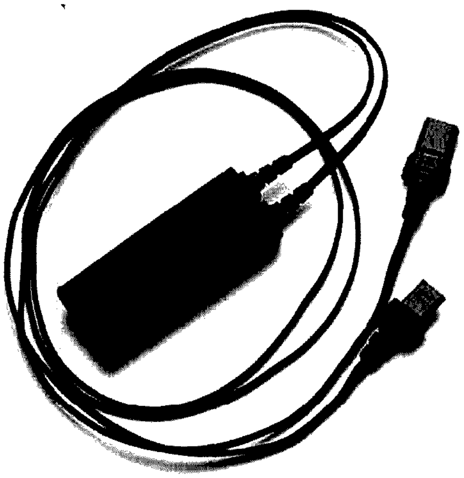

fig. 4. The HP-41 can be safely linked to the HP-IL by using the HP82160A module.

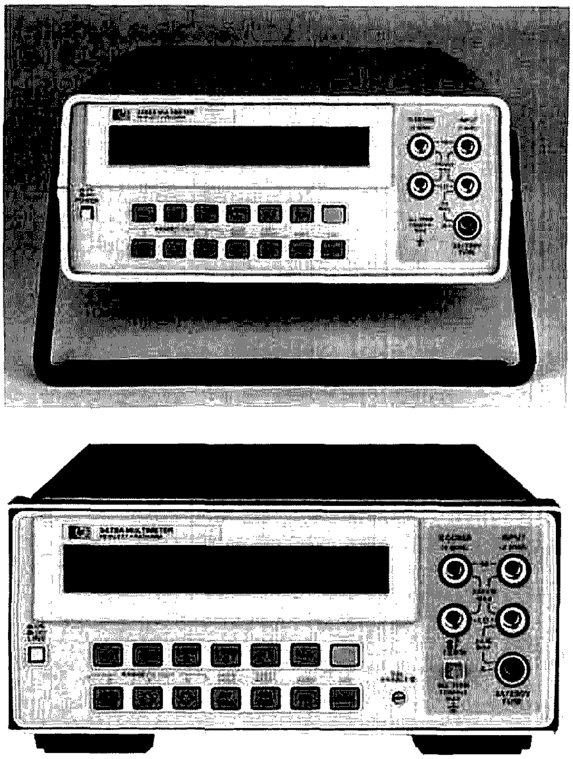

fig. 5. HP3468A HP-IL DMM and the HP3478A HP-IB DMM.

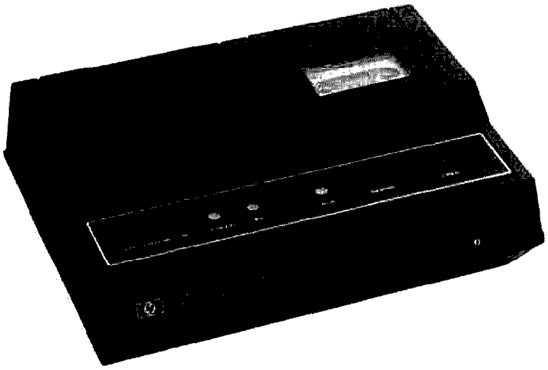

fig. 6. HP8216A HP-IL digital cassette drive.

ROMs, a card reader, a thermal printer, and a bar-code reader. Not surprisingly, HP-41 owners have yearned for a way to use these ports for communication with the outside world, but Hewlett-Packard has been reluctant to accommodate this desire, and with good reason: the unit's delicate CMOS circuitry can be easily damaged by improper connections made by overeager experimenters.

The HP-41, however, can be linked to the HP-IL via the HP82160A module (**fig. 4**), which plugs into any of the four ports mentioned above. The module receives

its power from the HP-41; two 28-inch (71 cm), two-wire cables extend from the module. Their ends are composed of two male and two female connectors. Cables can be up to 32.8 feet (10 meters) with simple stranded wire or up to 328 feet (100 meters) with twisted, shielded pair wire. Each HP-IL peripheral is in turn supplied with two connectors.

Applications. The unique ability of the HP-IL to poll its accessories allows the operator to simply enter a PRINT command; the network is then searched for the first available printer upon which any message in plain English may be printed. (This function would be ideal in a manufacturing facility in which relays had to be controlled, indicators read, and decisions made with this data that would allow plant operators to obtain essential information without delay.)

A DMM made just for the loop. The HP3468A DMM measures all five electrical parameters (AC and DC current and voltage, and resistance), is self-calibrating, and works in the loop. This 5 1/2-digit instrument costs $695 compared to its near-twin, the HP3478A, which costs $1300 (see **fig. 5**). While the 3478A does have added

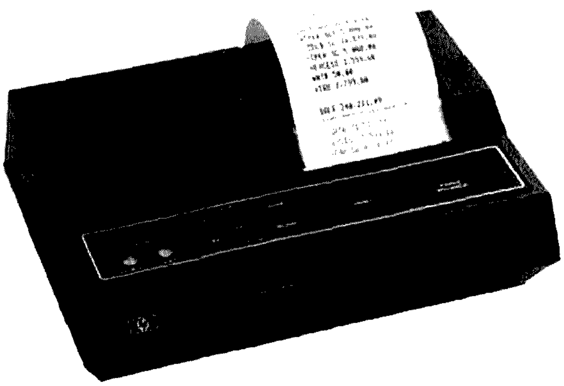

fig. 7. HP82162A HP-IL thermal printer.

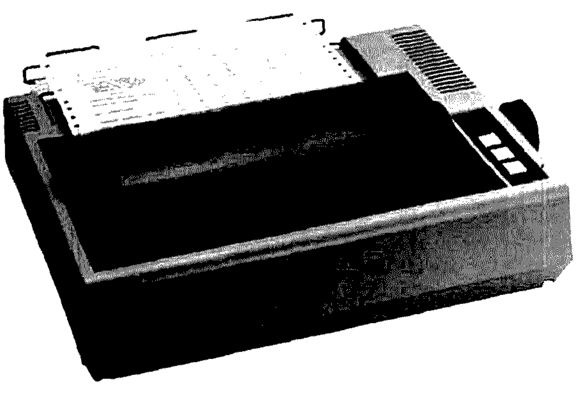

fig. 8. HP82905B impact printer.

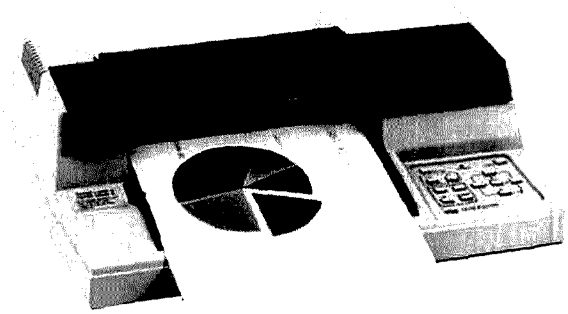

fig. 9. HP7470A graphics plotter.

fig. 10. HP82163A video interface.

fig. 11. HP82938A HP41-to-80 series computer interface.

RAM calibration constants and is characteristically much faster because it is an HP-IB, rather than an HP-IL, instrument, the HP-IL DMM, HP3468A, can be used with the HP-41 handheld calculator and other HP-IL instruments such as the HP8216A digital cassette drive and the HP82162A thermal printer (see **figs. 6** and **7**) to make a data logging system suitable for data from semiconduc-

tors or thermocouples. The HP41 also allows computation of complex math functions such as converting a voltage into dBs and then printing it out.

Other HP-IL devices include the HP82905B impact printer; the HP7470A graphics plotter; the HP82163A video interface, which permits 16 video lines of 32-characters to interface with a VHF TV; and the HP82938A interface, which allows the HP41 to link with the larger 80 series HP computers. Figs. 8 through 11 illustrate these devices.

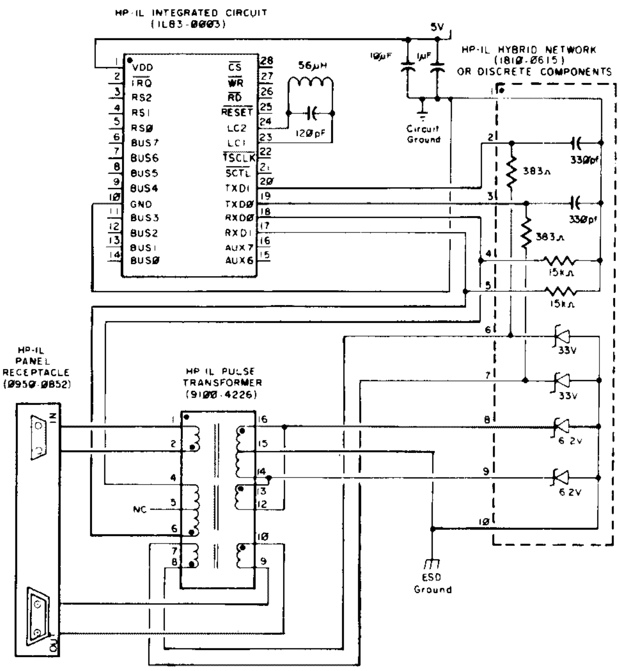

fig. 12. HP-IL interface connection diagram.

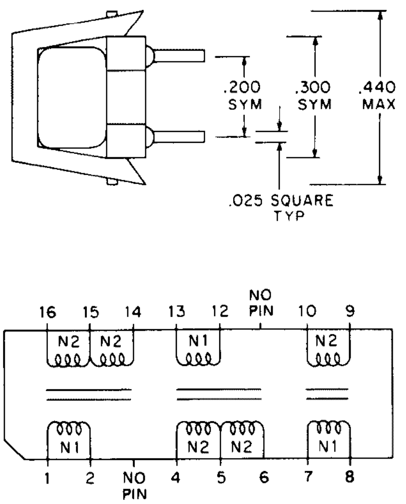

fig. 13. HP-IL transformer set (16-pin DIPheader).

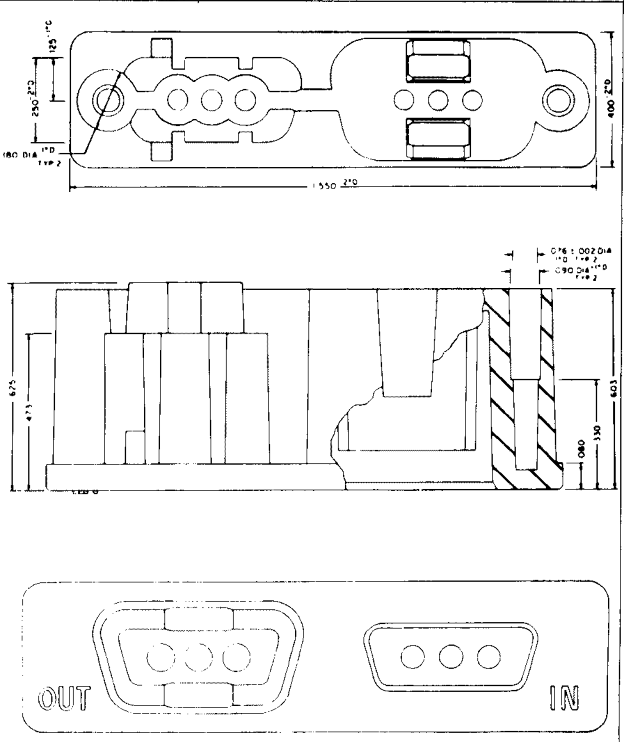

fig. 14. HP-IL panel receptacle.

The **HP-IL interface/prototyping kit.** This kit, the HP82166C, costs $395 and consists of two converters, a test board, cabling, and excellent documentation which allows prototyping of loop-compatible products. Its key components consist of the HP-IL interface connection, the HP-IL transformer set, and the HP-IL panel receptacle (see **figs. 12** through **14**).

The HP-IL IC appears to most standard microprocessors as eight memory or I/O locations. Reading and writing into its registers results in such actions as transferring data or a whole message and setting service request bits. This IC also handles all aspects of the loop's time-critical protocol. The ESD beside the ground symbol in **fig. 12** stands for *e*lectro *s*tatic *d*ischarge and is there for protection of the IC's delicate micropower CMOS circuitry. The transformer set provides for electrical isolation, impedance matching, and voltage level conversion, as previously mentioned, and the panel receptable set provides for foolproof mechanical connections and naturally proper direction to the flow of data.

reference

1. Vaughn D. Martin, "The HP-IB Greatly Simplified," *ham radio*, March, 1984, page 65.

ham radio

the branch-line hybrid: part 1

Applications include power splitting/combining, impedance transformation and attenuation

Connect four quarter-wave transmission lines end-to-end, add four RF connectors, and what do you have? A hybrid branch-line coupler, a simple device that cannot only divide power from a signal source, but also maintain isolation between the two output ports. And because the hybrid coupler is reciprocal, it can combine power from two equal phase and amplitude sources while maintaining isolation between the two inputs. If two amplifiers are paralleled together with a branch-line coupler, any input or output mismatch from one amplifier will not be reflected into the other.

The hybrid branch-line coupler consists of four transmission lines, two connected in series with the transmission path and two connected in shunt or "branching" the transmission line. Generally, one can expect a 0.5-dB maximum coupling imbalance, 14-dB return loss (SWR ≤ 1.5:1), and 15-dB isolation over a 20 percent bandwidth from a simple transmission line branch-line coupler. (These values can be improved as more sections are added, but the intermediate branching impedances become very difficult to realize.)

The single-section branch line coupler may be readily constructed using printed microstrip transmission lines and easily reproduced from a single etching mask at minimal cost. Experimental results are presented here to convince the Amateur that one needs only to compute the physical dimensions of the lines, etch the printed circuit board,[1] and cut the coaxial cable, in order to fully expect the coupler to function reasonably well over the Amateur band of interest. The lumped-constant version of the branch-line coupler is easily adjusted using a grid-dip meter.

The principal advantage of choosing the branch-line hybrid for power division and combining is the good input match and the amplifier output intermod improvement. Applications of the hybrid that illustrate these advantages are included at the end of this article.

derivation of coupler

When the power entering a network splits evenly between the output ports, the device is called a hybrid; if it divides in *any* other ratio, it is called a directional coupler. In a 20-dB directional coupler, for example, 99 percent of the power flows through one path while 1 percent of the input power appears at the other output port. The branch-line hybrid (**fig. 1**),

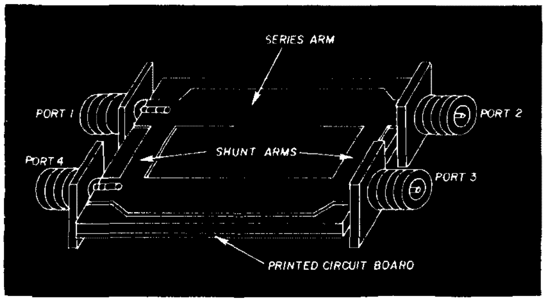

fig. 1. The branch-line hybrid can be formed using two series and two shunt microstrip transmission lines on a printed circuit board.

is composed of four quarter-wave transmission lines: two series arms and two shunt arms. Power input to port 1 is equally split between ports 2 and 3. Any mismatch power at either of these output ports appears

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at port 4. The output can be explained through use of either matrix manipulation or by sleight-of-hand. We shall use the latter.

The hybrid branch-line coupler, (**fig. 2**), is constructed using a series transmission arm, 1-2, and a shunt transmission arm, 2-3. Power input at port 1 travels along series transmission line 1-2. At point 2 the power is equally split between a 50-ohm load, placed at port 2 and shunt transmission line 2-3.

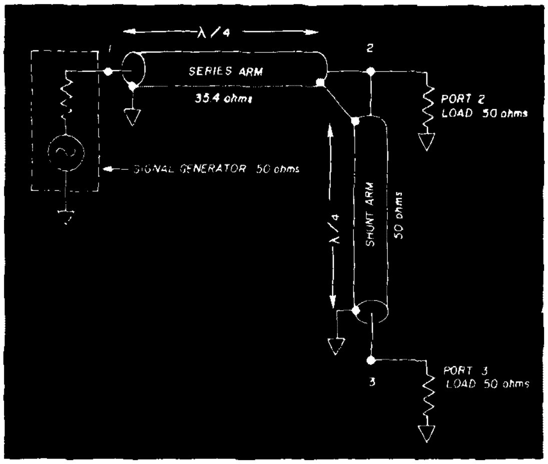


fig. 2. A simple equal-split power divider is formed using a series and a shunt transmission line.

Transmission line 2-3 is also 50 ohms, resulting in the parallel combination of the load and the input impedance to the shunt transmission line 2-3 of 25 ohms. This impedance must be transformed to present 50 ohms at input port 1. This is accomplished using series quarter-wave transmission line 1-2. The characteristic impedance of this quarter-wave matching section is equal to the square root of the product of the source and load impedances:

$$Z\,series\,arm = \sqrt{50\,ohms \cdot 25\,ohms} = 35.4\,ohms \quad (1)$$

The second output terminal, port 3, is also terminated with 50 ohms. This effectively terminates transmission line 2-3 in its characteristic impedance.

Thus we have formed a power divider using a series and shunt transmission line; but we have not, however, formed a true hybrid. Output ports 2 and 3 are isolated from each other by only 6 dB. Transmission lines 3-4 and 4-1 must be added, (**fig. 3**), to provide isolation between the output ports. The reflected wave due to any mismatch at port 2 is indicated as a dashed line traveling in two paths toward port 3. The counterclockwise path, however, is one half-wave longer than the clockwise path. Therefore, the

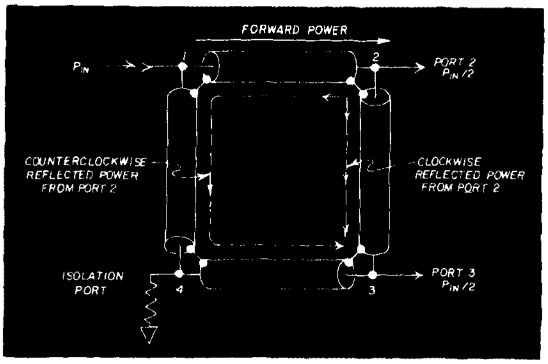


fig. 3. Reflected waves (dashed lines) from a mismatch at output port 2 cancel at output port 3 to provide isolation.

two reflected signals cancel at port 3 and port 3 is isolated from any mismatch at port 2. If the impedance mismatch occurs at port 3, the energy in the two paths likewise cancel at port 2.

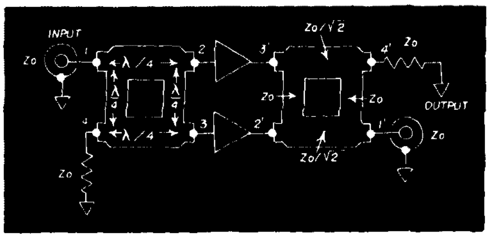


fig. 4. A 3-dB branch-line hybrid is used to parallel two amplifiers while maintaining isolation between them.

Isolation is especially necessary when combining two amplifiers in parallel, as shown in **fig. 4**. The output ports for the branch-line hybrid are 90 degrees apart, hence it falls into the category of quadrature hybrids. The output power at port 2 is delayed 90 degrees from the input power at port 1. The output power at port 3 undergoes an additional 90 degrees of phase delay. When recombining the power from the amplifiers, this must be considered by proper phasing.

The theoretical response of the single-section branch-line hybrid (**fig. 5**) is computed both for a 35.4-ohm and a 37.5-ohm series transmission arm. (The value of 37.5 ohms for the series transmission line was chosen because it can be simulated using two 75-ohm cables in parallel.) The 75-ohm coaxial cable is readily available from CATV sources. The graph *of input return loss* is an expression of the

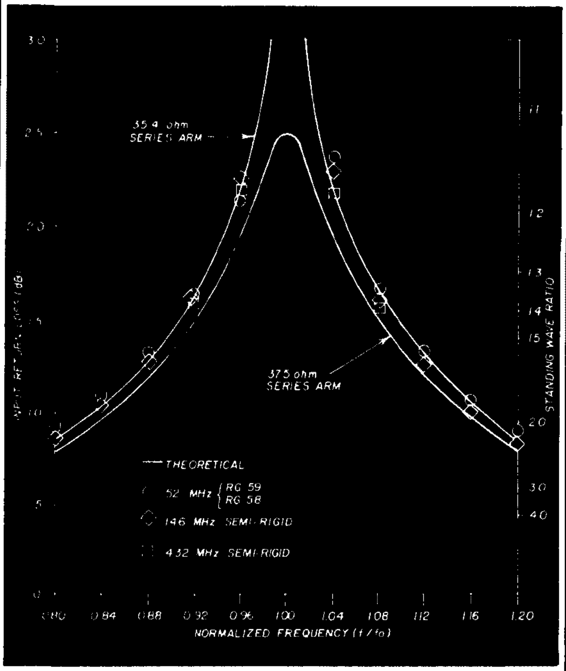


fig. 5. Experimental results for branch-line hybrids formed with coaxial cable closely follow the theoretical (solid line).

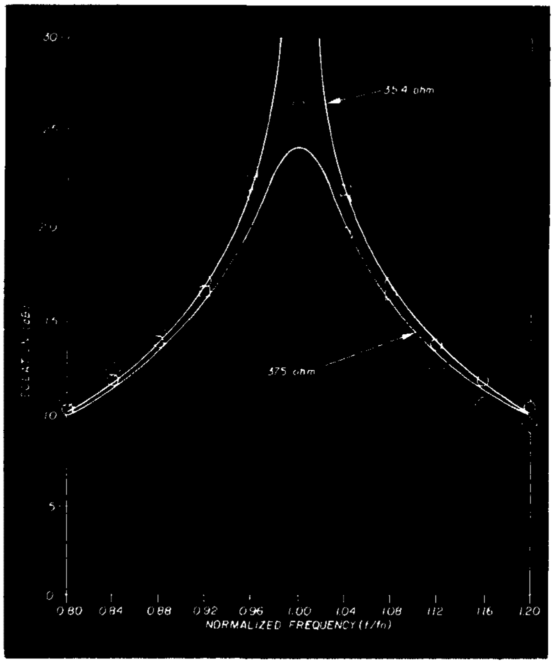


fig. 5B.

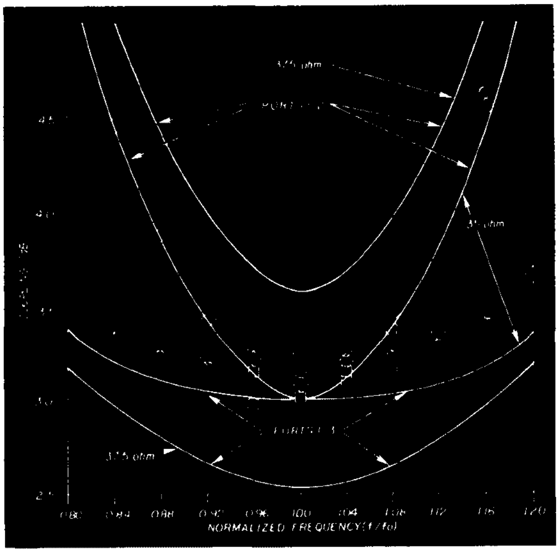


fig. 5C.

power ratio of the input power to reflected power in decibels. To convert return loss to the familiar SWR, the following equation is used:

$$SWR = \frac{10^{RL/20} + 1}{10^{RL/20} - 1} \qquad (2)$$

where RL is the return loss.

To convert from SWR back to input return loss:

$$return\ loss = 20 \log_{10} \left[\frac{SWR + 1}{SWR - 1}\right] \qquad (3)$$

Expressing the value of the input match in terms of return loss is becoming more popular. This is the ratio indicated on a directional wattmeter expressed in dBs. A value of 6 dB input return loss ($SWR = 3{:}1$) means that one-fourth of the input power is reflected. The value of theoretical coupling loss for a hybrid assumes no insertion loss. Hybrids typically display an insertion loss of 0.1 to 0.2 dB due to transmission line loss or low-Q lumped-constant elements (loss). A figure of 3.01 dB expresses an equal power division between ports 2 and 3 and does not mean that the hybrid dissipates half the power. The value of isolation is measured between ports 2 and 3. A signal applied to port 2 will be reduced by the value of the isolation before arriving at port 3. This is equally true for an input signal reflected at output port 2 appearing at port 3. Because the hybrid is symmetrical, this is also the isolation between ports 1 and 4.

The models constructed using flexible coaxial cable employed two 75-ohm quarter-wave sections in parallel to simulate each series arm (**fig. 6**). The parallel impedance is 37.5 ohms, which is close to the desired 35.4 ohms; the difference in response is shown in **fig. 5**. Each shunt arm was formed with

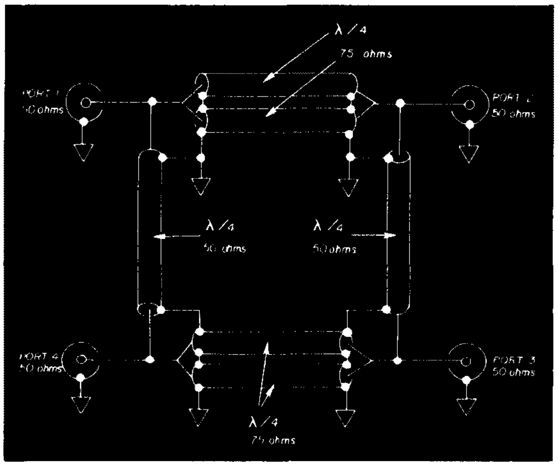


fig. 6. A 50-ohm branch-line hybrid may be formed with readily available 50-ohm and 75-ohm coaxial cable.

common 50-ohm coaxial cable. Models constructed with semi-rigid coaxial cable relied on two 70-ohm lines in parallel, which yielded results very close to the theoretical value of 35.4 ohms.

The physical length, L, of each quarter-wavelength transmission line may be calculated using:

$$L = \frac{c \cdot V_p}{4f} \qquad \textbf{(4)}$$

where c = speed of light (3×10^8 m/sec) or 9.84×10^8 ft/sec
f = frequency (Hz)
V_p = relative velocity of propagation = $1/\sqrt{e}$, where e is the relative dielectric constant

dielectric	V_p
solid polyethylene	0.66
foam polyethylene	0.78
solid teflon	0.69

The value of L, (**table 1**), is calculated in the same units as the speed of light constant chosen.

The coaxial cable may be trimmed exactly by measuring its resonance with a grid-dip meter after cutting the cable slightly longer than the length computed above and preparing the ends. The center conductor is shorted to the braid at one end only, and the grid-dip meter is positioned close to the loop formed at the shorted end. The shorted quarter-wavelength transmission line acts as a parallel-resonant circuit.

If the branch-line hybrid is redrawn in a circular manner (**fig. 7**), rather than in the square representation, it resembles the familiar hybrid ring or "rat race."[2] The hybrid ring, however, uses three sections of quarter-wavelength line joined together by a three-quarter-wavelength section.

next month:

In part 2, branch-line hybrids that use lumped constant components for operation below 100 MHz are described. Additional applications discussed include impedance transformers and pin-diode attenuators.

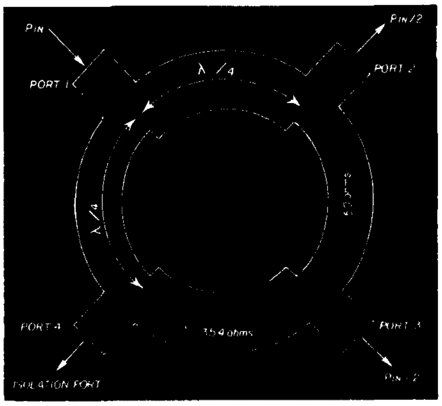


fig. 7. Circular form of the branch-line hybrid resembles the hybrid ring.

table 1. Cable length for various dielectrics.

frequency (MHz)	solid polyethylene (0.66)	foam polyethylene (0.78)	solid teflon (0.69)
3.75	43 ft 3 in	51 ft 1 in	45 ft 2 in
7.15	22 ft 8 in	26 ft 9 in	23 ft 8 in
14.20	11 ft 5 in	13 ft 6 in	11 ft 11 in
21.25	7 ft 8 in	9 ft 0 in	7 ft 11 in
28.50	5 ft 8 in	6 ft 9 in	5 ft 11 in
52.00	37.4 in	44.2 in	39.1 in
146.00	13.3 in	15.7 in	13.9 in
222.00	8.7 in	10.3 in	9.2 in
435.00	4.47 in	5.29 in	4.67 in
1250.00	1.56 in	1.84 in	1.63 in

ham radio

book and product REVIEWS

action monitor

JC Labs of Wales, Wisconsin, has introduced the Action Monitor, a voice-operated switch (VOX) that will connect between any ordinary cassette or other type tape recorder and a receiver.

Hookup is simple. You feed the audio from the receiver to the Action Monitor. The audio then goes two ways: one to a speaker, the other to the mike input on the tape recorder. You also connect the Action Monitor to the tape recorder's on-off switch. As soon as the Action Monitor "hears" a transmission, it activates the tape recorder via the on-off switch; when the transmission is completed, the unit will continue taping for several seconds to record any replies and then the Action Monitor shuts off the tape recorder until another message is received.

The best thing about the Action Monitor is what it does: eliminate the need for a high priced, special-purpose type recorder. All you need is an ordinary cassette recorder, or any other type recorder, a receiver and the Action Monitor, and you're ready to record any service you wish. Whether you want a permanent record or just want to know what went on while you were asleep or away from home, the Action Monitor will provide a taped record of what transpired. Some suggested uses include recording from the business bands, police and fire calls, Amateur Radio repeaters, aircraft band, even CB. (You must remember that all transmissions are protected by the Secrecy in Communications Act.)

The next best thing about the Action Monitor is its price: only $39.95 plus $2 for shipping in the U.S., $3 for shipping to Canada.

The lightweight unit comes in a sturdy black plastic case measuring 2-3/4 × 4 × 1-1/2 inches. It operates from a 9-volt battery or optional AC power supply ($8.95).

For complete information, contact JC Labs, Inc., P.O. Box 183, Wales, Wisconsin 53183.

— N1ACH

Circle #301 on Reader Service Card.

non-iambic paddle

Bencher has just released a new single lever, non-iambic paddle for hams like me who never were able to master the intracacies of iambic keying.

Bencher paddles use stainless steel bearings, stainless fasteners and a stainless steel, lifetime spring. Dot-dash contacts are gold-plated silver to ensure positive contact. The heavy leaded steel base is designed to eliminate paddle "creep" while being used. All in all, the paddles are built for a lifetime of use and the quality of their construction reflects that fact.

The non-iambic paddle is manufactured to the same standards as Bencher's iambic paddle. When the review unit was received, I eagerly opened it up and wired it to my keyer. I felt immediately at home with the new paddle.

You can adjust two basic parameters on the Bencher key. The first is spring tension, which is adjusted by either tightening or loosening a ferruled nut at the back of the key. I favor light spring tension, so I set the spring on the review unit quite loosely. The dot and dash contacts are adjusted by loosening a lock-screw and then moving them either closer or farther away with an Allen wrench. In adjusting my iambic key I misplaced my Allen wrench and wound up searching for it any time adjustments were necessary. Fortunately Bencher has added a Fahrnstock clip underneath the non-iambic paddle to eliminate losing the Allen wrench altogether.

I haven't used the new key in a contest yet, so I can't report on how well it works or stays put when I get excited. I would expect that it will work rather well.

The Bencher non-iambic key comes in three models: with black base at $46.95, chrome base at $59.95, and a gold plated presentation model at $150.

For more information, contact Bencher at 333 West Lake Street, Chicago, Illinois 60606.

Circle #302 on Reader Service Card.

N1ACH

Novice Class Amateur Radio Operator Test Guide

The name Bash evokes mixed feelings in the Amateur community. Some love him, others dislike him intensely. His latest book, however, should be favorably received by instructors and prospective Novices alike.

Since the Novice exam is written and administered by the examiner, the controversial question-and-answer-format used in Bash books for General, Advanced and Extra Class licenses will not work. Instead, Bash has written an easy-to-read, complete, informative, but extremely informal Novice licensing guide.

Bash starts off with the basics of getting a license, Amateur Radio rules and regulations, and electronics. As you progress through the book, Bash explains many of the mysteries of radio and electronics theory without getting needlessly complex. One problem with other study guides is that they sometimes tell you much more than you need to know. Bash cuts away at all extraneous information and gives you exactly what you need to be capable of passing a test based upon the FCC exam syllabus. For those who are looking to go beyond the scope of a basic study guide, Bash makes extensive reference to *Amateur Radio Theory and Practice* by Robert Shrader, W6BNB. This handy cross-reference virtually ensures that a prospective Novice can get the exact answer that is needed.

Bash also has a 34-question Q&A section that will help the reader test his or her learning prior to taking the exam. And so that the prospective Novice can be fully prepared for the exam, Bash has included a complete copy of the official FCC Novice exam study guide.

Finally, Bash has gone to the trouble of filling the book with handy tips and hints that will help answer many of the often-asked questions about radio operation.

For prospective Novices planning to study independently or in organized classes, Bash's new Novice study guide, priced at $9.95, will be helpful. (Copies are available from Ham Radio's Bookstore; add $1.00 for postage and handling.)

For more information, contact Bash Educational Services, P.O. Box 2115, San Leandro, California 94577.

Circle #303 on Reader Service Card.

N1ACH

NEW products

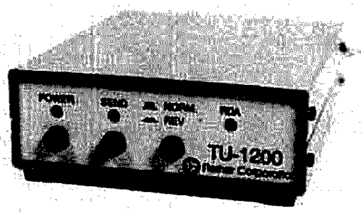
UHF/VHF RTTY terminal

The new TU-1200 UHF/VHF RTTY terminal unit from Flesher receives all Baudot and ASCII rates to 1200 baud and uses Bell 202 standard tones (1200 Hz and 2200 Hz) useful in many applications, including RTTY repeater systems. It provides TTL and RS-232C compatible I/O and includes transmitter PTT output for complete remote control. The TU-1200 also provides AFS output and RDA (received data available).

Only three push-button switches are required for operation: POWER, SEND, and NORMAL/REVERSE SHIFT. Three LED indicators show POWER, SEND, and RDA status. The TU-1200 is constructed with a quality all-metal case to assure proper RF protection. Its small

size (5 1/8"W × 1 3/4"H x 6"L) and rear panel DB-25 I/O connector make installation and use simple.

The TU-1200 is available either wired ($129.95) or in kit form ($99.95) and comes complete with a mating DB-25 I/O plug, power sup-



ply and an easy-to-understand step-by-step OPERATOR/ASSEMBLY manual.

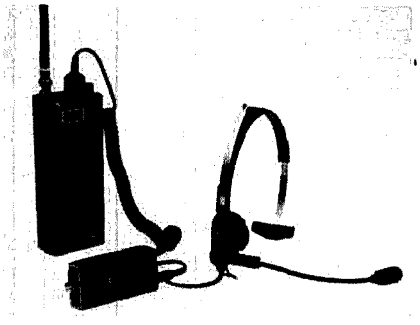
For more information, contact Flesher Corporation, P.O. Box 976, Topeka, Kansas 66601.

Circle #304 on Reader Service Card.

ICOM accessories

The new ICOM IC-HS10 Headset and IC-HS10SB PTT Switchbox can be used with all ICOM handheld transceivers: the IC-2A and 2AT; IC-3A and 3AT; IC-4A and 4AT; IC-02A and 02AT; and IC-04A and 04AT. The lightweight headset includes the following features: crystal-clear reception, pivoting microphone, and adjustable boom; it folds up for storage and adjusts for comfortable fit.

The switchbox measures 3 × 1.5 × 0.75 inches and features a belt clip, transmit-receive switching control, mic gain control, and a molded plastic connector for speaker/mic connection to handheld.



The IC-HS10 Headset and IC-HS10SB PTT Switchbox are available immediately and may be purchased separately for $19.50 each or $39.00 for the set.

For more details, contact ICOM, 2112 116th Ave. N.E., Bellevue, Washington 98004.

Circle #305 on Reader Service Card.

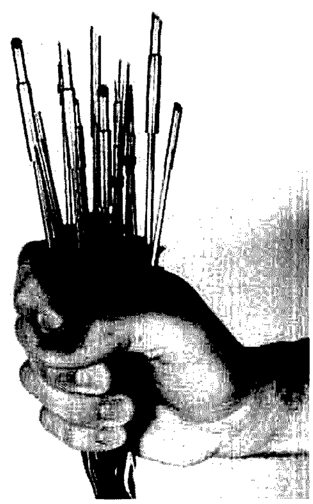
miniature soldering system

With Wahl's new assortment of 15 miniature soldering irons and 23 tips, a user can choose from over 270 different soldering combinations to match precise soldering needs. The 15 Oryx® miniature irons, all weighing 1/2 ounce, or less, are available in temperatures ranging from 575 to 850 degrees F; from 5 to 25 watts; and from 4.5 to 24 volts. Their compact size and precise temperature control make them a good choice for soldering heat-sensitive components.

The irons can be combined with any of 23 tips ranging in size from 1/25 to 3/32 inch in several choices of configuration. Tip construction is nickel-plated or iron-plated copper for most applications, with solid nickel, gold end, and bare copper alloy (NASA) tips available for special requirements. Tip changes are easy, with no tools required. Cooled tips simply slide off and on.

For further information, contact Wahl Clipper Corporation, Sterling, Illinois 61081.

Circle #306 on Reader Service Card.



NEW products

new antenna rotator for blind hams

Telex/Hy-Gain has introduced the HAM-SP rotator specially designed for visually impaired Amateur Radio operators.

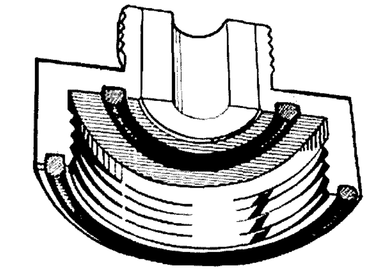
The control unit functions are marked in both Braille and conventional lettering. The unit also emits a high frequency tone to indicate rotator action. Since the brake release as well as delayed brake engagement is automatic, operation of the rotator is a simple one-hand, one-touch operation.

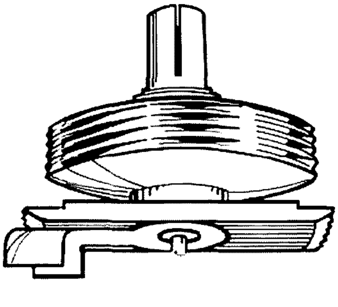
When mounted inside a tower, the new HAM-SP rotator is designed to operate large antenna arrays of up to 15 square feet (1.4 m^2) wind load area. The HAM-SP (catalog No. 307) carries a suggested list price of $337 and is available at Amateur Radio dealers.

For further information, contact Telex Communications, 9600 Aldrich Ave. So., Minneapolis, Minnesota 55420.

mounting kits

The Larsen PO-K mounting kit features SO-239 style mounting hardware that can be installed entirely from outside any vehicle. The PO-series mount looks like the SO-239 connector





(sometimes called a female UHF connector), but unlike the SO-239 can be installed from outside. Its two "O" rings and one gasket offer superior moisture sealing. The PO-K includes complete mounting kit and coax; the PO-B contains mounting hardware only.

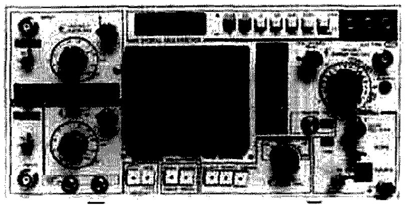
For further information, contact Larsen Electronics, P.O. Box 1799, Vancouver, Washington 98668.

Circle #307 on Reader Service Card.

digital storage oscilloscope

The Model MS-3020 Multi-Scope from North American SOAR employs features never before available in a single compact package, including a 15-MHz triple trace real time oscilloscope with 3 1/2 inch flat face internal graticule CRT; variable trigger delay plus single sweep; a built-in "quick tracer" type component tester for cold circuit, individual component or full board circuit evaluation; a five-function 3 1/2 digit LED display DMM that can be operated independently or simultaneously in the same circuit as the oscilloscope; a digital storage section with 1024 words of memory; and rear panel BNC connectors for pen recorder output of X-Y sync of the stored Channel 1 waveform.

The MS-3020 is housed in a small, easily transportable case with U bracket handle measuring 10 inches wide × 5 inches high × 12 1/4 inches deep. It weighs approximately 13 1/2 pounds and is priced at $1995.



For more information, contact North American SOAR Corporation, 1126 Cornell Avenue, Cherry Hill, New Jersey 08002.

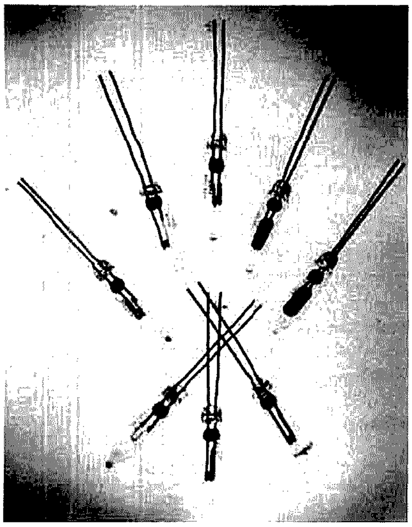
Circle #308 on Reader Service Card.

thermo protectors

A series of thermo-protectors designed to provide efficient, low-cost protection for electrical equipment, are included in the EGG semiconductor replacement line available from the Distributor & Special Markets Division of Philips ECG, Inc.

Designated ECG 202, 204, 205, the units are intended for replacement use in small fractional horsepower motors and products such as audio speakers, transformers, HID lamps, automotive antenna motors, CATV amplifiers, and television circuits.

Each thermo-protector contains a prestressed bimetal blade; fault conditons producing excessive current or heat will cause the bimetal to snap the contacts open. Encased in a strong, shock-



resistant hermetically sealed glass cartridge, the unit can withstand external pressures exceeding 5,000 PSI.

The entire ECG semiconductor replacement line, including more than 3200 solid-state devices, is available from authorized distributors of Philips ECG components.

For more information, contact Philips ECG, Inc., 100 First Avenue, Waltham, Massachusetts 02254.

Circle #309 on Reader Service Card.

Worldwide Sunrise/ Sunset Tables

Noted 80-meter DXer and author John Devoldere, ON4UN, has just released the latest edition of the *World Wide Sunrise/Sunset Tables.'*

As sunspot activity declines, more and more DX'ers are turning to 160, 80, and 40 meters and finding that "gray line" propagation is an integral part of successful DX'ing on those bands.

ON4UN's book provides sunrise and sunset times for 502 geographical locations in all DXCC countries and includes 100 listings in the U.S.A. Also included are fairly detailed instructions so the reader can accurately utilize the information in the book. For example: Is there a long path between San Francisco and Budapest on January 1?

In looking at both San Francisco and Budapest, we find:

San Francisco	Sunrise West 1524	Sunset West 0058
Budapest	Sunrise East 0632	Sunset East 1503

NEW products

For long path conditions to exist, you must be at or near a peak propagation path at sunrise with the station at the western end of the path or at sunset with the station at the eastern end of the path. With our example Sunrise West is later than Sunset East and there will be a 1/2 hour-long path opening between approximately 1500 and 1530 hours Z.

Devoldere also talks about crooked paths between two locations in the same hemisphere in the middle of local winter, such as exists between the U.S. and Japan during December and January. These paths do not follow the "great circle" paths and are skewed by passage through the Polar regions area.

This book is available directly from the author and comes with a personalized beam heading printout and an enlarged personalized Sunrise/Sunset Chart. Price is $10 postpaid (air mail).

For more information, contact the author directly at:

John Devoldere, ON4UN
215 Poelstraat
B9220 Merelbeke
Belgium

lower price for L/C meter

The Model 522 digital L/C meter and test fixture combination is being offered by Cambridge Technology at a new low price — $875.00.

This instrument combination allows measurement and sorting of inductances and capacitances at a basic accuracy of 0.25 percent, with dual measurement frequencies, and 0 to 10 VDC internal capacitor bias, for less than $1000.00. (The closest competitor to the Model 522 costs over 40 percent more.)

A built-in comparator allows simultaneous sorting for high and low tolerances, as well as dissipation limits. Measurement frequencies of 120 Hz or 1 kHz are automatically selected. The instrument is autoranging and a range-hold feature permits fast repetitive measurements.

The electrical contacts of the test fixture provide four-terminal measurement capability. A push button allows both electrical contacts to be held open simultaneously so that components with very fine leads can be inserted conveniently. A second push button allows each set of contacts to be individually opened for components with flying leads such as coils or transformers.

Other features include a rugged all-metal enclosure, tilt stand, and lead capacitance null. The instrument is protected against damage from charged capacitors and comes with a one-year warranty on materials and workmanship. A full refund within 30 days is available if the instrument is found unsatisfactory for any reason.

For more information contact Cambridge Technology, Inc., 2464 Massachusetts Avenue, Cambridge, Massachusetts 02140.

Circle #310 on Reader Service Card.

selective call controller

The SCC-1 Selective Call Controller, from Acquis Communications, Inc., was designed for use with mobile or base FM radio systems. Its features include a touchtone encoder; 10-number auto dialer with battery backup; a touchtone decoder with two programmable selective call codes for group or individual calls; an internal monitor speaker; LED displays for call status; and an accessory relay for control of external devices. The small size of the SCC-1 (3 × 4 × 2 inches) and flexible mounting hardware make it easy to adapt for either mobile or base station applications. A mobile mounting bracket and protective desk pads are included.

The SCC-1, priced at $325, is particularly useful for emergency groups such as RACES or radio clubs. The selective call feature with programmable call codes allows for group or private monitoring of radio communications.

For further information, contact Acquis Communications, Inc., 17192 Gillette Avenue, Irvine, California 92714.

Circle #311 on Reader Service Card.

code practice aid

The Noise Maker was originally conceived to assist those with problems hearing code as it is usually generated. We usually listen to code in the form of a keyed tone — usually pure — somewhere in the range from 500 to 1000 Hz. But because this kind of sound is troublesome for some listeners, W6NRW put together a white-noise generator with a keying circuit designed to enable fast, click-free keying.

His idea was to spread the energy over a wide audio spectrum, in the expectation that enough sound energy would be emitted within a person's hearing range to enable reception. Since the sound energy generated within any particular narrow frequency range is low, the listener will not be irritated by high levels of tones that may be troublesome. And because white noise is a more "natural" form of sound energy than a pure tone, practicing with the noise maker may cause less fatigue.

For more information on the Noise Maker ($12.95 plus $2.50 shipping and handling) and an optional electronic switch that allows white noise code practice using conventional oscillator tapes or a receiver ($7.00), contact Hildreth Engineering, P.O. Box 60003, Sunnyvale, California 94088.

Circle #312 on Reader Service Card.

model 234 encoder

SYT had introduced a new low-cost mobile DTMF encoder for autopatch and selective call radio systems. This compact device features a new high-speed, crystal controlled DTMF encoder circuit with a tactile feedback keypad, ATK, and interdigit hold timer. The 234 is housed in an attractive metal housing with a variable tilt mounting bracket. Ideal for small car installations, the 234 is the first budget-priced, quality mobile encoder available.

For details, contact SYT Corporation, 1220 Barranca Street, El Paso, Texas 79935.

Circle #313 on Reader Service Card.

equipment protection

A new 40-page catalog (No. 831) from Electronic Specialists presents their line of protective and interference control equipment. Protective devices available include equipment isolators, AC power line filter/suppressors, AC line voltage regulators and modem surge suppressors.

Descriptive sections outline specific communication problems and suggested solutions. Typical applications and uses are highlighted.

For further information, contact Electronic Specialists, Inc., 171 South Main Street, Natick, Massachusetts 01760.

Circle #314 on Reader Service Card.

flea market

RATES Noncommercial ads 10¢ per word; commercial ads 60¢ per word **both payable in advance.** No cash discounts or agency commissions allowed.

HAMFESTS Sponsored by non-profit organizations receive one free Flea Market ad (subject to our editing) on a space available basis only. Repeat insertions of hamfest ads pay the non-commercial rate.

COPY No special layout or arrangements available. Material should be typewritten or clearly printed (**not** all capitals) and must include full name and address. We reserve the right to reject unsuitable copy. **Ham Radio** cannot check each advertiser and thus cannot be held responsible for claims made. Liability for correctness of material limited to corrected ad in next available issue.

DEADLINE 15th of second preceding month.

SEND MATERIAL TO: Flea Market, Ham Radio, Greenville, N. H. 03048.

QSLs & RUBBER STAMPS — Top Quality! Card Samples and Stamp info — 50¢ — Ebbert Graphics 5R, Box 70. Westerville, Ohio 43081.

TRAVEL-PAK QSL KIT — Converts post cards, photos to QSLs. Stamp brings circular. Samco, Box 203-c, Wynantskill, New York 12198.

WANTED: Manual for Dumont oscillograph type 304AR. WU4C, Leon, Rt. #6, Box 751A, Macon, GA 31201.

COUNSELORS: Connecticut brother-sister camp. Completely equipped with ham radio station. Program includes electronics, kit building, code and communications. June 25 - August 22. Send resume: Lloyd Albin (N2DMQ) Ken-Mont and Ken-Wood Camps, 2 Spencer Place, Scarsdale, NY 10583.

ATTENTION DXers! Get Russian cards faster. All local Soviet QSL Bureau addresses $4.00. Ed Kritsky, KA1MXO, PO Box 715, Brooklyn, NY 11230.

NEED TO CONTACT James Navarchi concerning Yaesu gear. C.T. Huth, 146 Schonhardt St., Tiffin, OH 44883.

ATTENTION C-64 USERS. Don't buy a logging program until you've read our fact sheet. For free information write to Crumtronics, PO Box 6187, Ft. Wayne, IN 46896.

Foreign Subscription Agents for Ham Radio Magazine

Ham Radio Austria Karin Ueber Postfach 2454 D-7850 Loerrach West Germany	Ham Radio Holland Postbus 413 NL-7800 Ar Emmen Holland
Ham Radio Belgium Stereohouse Brusselsesteenweg 416 B-9218 Gent *Belgium*	*Ham Radio Italy* Via Pordenone 17 I-20132 Milano Italy
Ham Radio Canada Box 400, Goderich *Ontario*, Canada N7A 4C7	Ham Radio Switzerland Karin Ueber Postfach 2454 D-7850 Loerrach West Germany
Ham Radio Europe Box 2084 S-194 02 Upplands Vasby Sweden	
Ham Radio France *SM Electronic* 20 bis, Ave des Clarions F-89000 Auxerre France	Ham Radio UK *P O Box 63, Harrow* Middlesex HA3 6HS England
Ham Radio Germany Karin Ueber Postfach 2454 *D-7850 Loerrach* West Germany	Holland Radio 143 Greenway Greenside, Johannesburg Republic of South Africa

FOREIGN PAPER MONEY wanted for my hobby. Old and new. Will accept free or will buy or trade. Buddy Hincke, 1854 East Bay Drive, North Bend, Oregon 97459. WA6LFJ.

BEAM HEADINGS, computer calculated for your QTH, $6.00. Commodore-64 programs for hams. Huff, Box 1112, Springfield, IL 62705.

120' GUYED TOWER. Extremely strong $700. 20' sections $150. Tim Colbert, 13609 Colony, Burton, Ohio 44021.

FREE, FREE gift. Interested in Amateur Radio, computers, video. Large SASE pse and mention Ham Radio ad. Free gift to all. Narwid Electronics, 61 Bellot Rd., Ringwood, NJ 07456.

VLF-LF preamps, coupler, Loran-C boards. SASE. Burhans Electronics, 161 Grosvenor St., Athens, Ohio 45701.

WANTED: MFJ-1224 computer interface, RS232 cable, software for H89A. Zenith trans-oceanic model 7000 radio. W4BUZ, Don Harris, 1703 Verdun Dr., Rt. 11, Greensboro, NC 27410.

ROHN TOWERS — Wholesale direct to users. All products available. Write or call for price list. Also we are wholesale distributors for Antenna Specialists, Regency, and Hy-Gain. Hill Radio, PO Box 1405, 2503 G.E. Road, Bloomington, IL 61701-0887. (309) 663-2141.

ELECTRON TUBES: Receiving, transmitting, microwave . . . all types available. Large stock. Next day delivery most cases. Daily Electronics, 14126 Willow Lane, Westminster, CA 92683. (714) 894-1368.

WANTED: Technical, computer and broadcast books by Harold E. Ennes. C.J. Klaue, 9871 Martinique Drive. Miami, FL 33189.

RUBBER STAMPS: 3 lines $4.50 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

"THE SWAP LIST" has bargains galore. Subscribe now! 6 months for $4.00; 1 year only $6.50. The Swap List, Box 988-H, Evergreen, CO 80439.

CHASSIS and cabinet kits. SASE K3IWK.

SAN ANTONIO, TEXAS QTH for sale. 4-2-2 with 70 foot tower, etc. W8CM, (512) 684-6129.

CABLE CONVERTERS, decoders. Catalog $1 refundable. APS, POB 263 HR, Newport, RI 02840.

VHF + TRADING POST. Buy, sell, trade your VHF, UHF and microwave equipment and accessories. SASE for sample. 1420 East Sweet Avenue, Bismarck, ND 58501.

WANTED: Cash paid for used speed radar equipment. Write or call: Brian R. Esterman, PO Box 8141, Northfield, Illinois 60093. (312) 251-8901.

WANTED, MILITARY SURPLUS RADIOS. We need Collins 618T, ARC-72, ARC-94, ARC-102, RT-712/ARC-105, ARC-114, ARC-115, ARC-116, RT-823/ARC-131, or FM-622, RT-857/ARC-134 or Wilcox 807A, ARC-159, RT-1167 or RT-1168/ARC-164, RT-1299/ARC-186, RT-859/APX-72, APX-76, ARN-82, ARN-84, ARN-89, APN-153, APN-155, APN-171, MRC-95, 718F-1/2, HF-105, Collins antenna couplers 490T-1, 490T-2, 490T-9, CU-1658A/ARC, CU-1669/GRC, 490B-1, 690D-1, CU-1239/ARC-105. Top dollar paid or trade for new Amateur gear. Write or phone Bill Slep, 704-524-7519, *Slep Electronics Company, Hwy. 441,* Otto, NC 28763.

RECONDITIONED TEST EQUIPMENT $1.00 for catalog. Walter, 2697 Nickel, San Pablo, CA 94806.

FOX-TANGO Newsletters — Since 1972, the prime source of modifications, improvements, and repair of Yaesu gear, free to Club members. Calendar year dues still only $8 U.S., $9 Canada, $12 elsewhere. Includes five year cumulative index by model numbers, or send $1 for index and sample Newsletter. Fox Tango Club, Box 15944, W. Palm Beach, FL 33416.

RTTY-EXCLUSIVELY for the Amateur Teleprinter. One year $7.00. Beginners **RTTY** Handbook $8.00 includes journal index. P.O. Box RY, Cardiff, CA 92007.

OSCILLOSCOPE, 3 inch CRT Dumont, Model 292. Works fine. Unmarked. $40.00. KS6N. (209) 478-9092.

IMRA International Mission Radio Assn. helps missioners — equipment loaned; weekday net, 14.280 MHz, 2-3 PM Eastern. Br. Frey, 1 Pryer Manor Rd., Larchmont, NY 10538.

4-400A tubes. Pull-outs but deliver 600W. Three at $25.00 each. KS6N. (209) 478-9092.

"HAMS FOR CHRIST." Reach other Hams with a gospel tract sure to please. Clyde Stanfield, WA6HEG, 1570 N. Albright, Upland, CA 91786.

FOR SALE. Exceptional 28ASR Teletype. 3-speed gearshift typing unit with reperforator, dome mounted reperforator with 3-speed gearshift, 2-amp regulated loop supply, separate here-is unit 60 wpm. Very clean. $375 plus shipping, price negotiable with removal of accessories. Also Model 28TD and keyboard reperforator, both 60 wpm. Bill K3PGB, 1257 Wunderland Road, Roslyn, PA 19001.

TENNATEST — *Antenna noise bridge — out-performs others,* accurate, costs less, satisfaction guaranteed. Send stamp for details, W8URR, 1025 Wildwood Road, Quincy, MI 49082.

WANTED: Early Hallicrafter "Skyriders" and "Super Skyriders" with silver panels, also "Skyrider Commercial", early transmitters such as HT-1, HT-2, HT-8, and other Hallicrafter gear, parts, accessories, manuals. Chuck Dachis, WD5EOG, The Hallicrafter Collector, 4500 Russell Drive, Austin, Texas 78745.

VERY in-ter-est-ing! Next 4 issues $2. Ham Trader "Yellow Sheets", POB356, Wheaton, IL 60189.

ANNIE's EASY. Analyze dipoles, slopers, verticals, inverted-vees and arrays; any orientation, position, phasing, weight or combination with Annie Antenna Analysis Software. Include REAL GROUND (conductivity, dielectric constant). Superb hi-res plotting. Annie's incredibly friendly and with 100% machine *language, she's FAST! For Apple][+ (48K or 64K) or][e,* DOS3.3, $49.95 + $2.00 postage, NY add sales tax. Include full name and call. S.A.S.E. for info. Commercial, library, etc., call for quote (315) 622-3641. Sonnet Software, Dept. HR, 4397 Luna Course, Liverpool, NY 10388.

RF TECHNICIAN: Looking for an energetic, self-starter with bench experience. Experience with RF, electronics, and computers important. Ground floor opportunity with medical device-manufacturer. International travel required. Potential limitless. Domestic base in northern New Jersey. Send resume and salary requirements to: TechTron, Dept. PTS, 101 Old Short Hills Road, Suite 215, West Orange, New Jersey 07052.

Coming Events

ACTIVITIES

"Places to go..."

ARIZONA: The Cochise Amateur Radio Association is sponsoring a Hamfest, May 4, 5 and 6 in St. David. Tours are planned to Tombstone, the Bisbee Lavendar Pit and other places of interest. There will be a flea market and tailgaters are welcome. Talk in on 16/76 and 52 simplex. For details contact C.A.R.A., PO Box 1855, Sierra Vista, AZ 85636, attention KB7HB.

ARKANSAS: The Northwest Arkansas ARC will hold its 4th annual Hamfest/Swapmeet, Saturday, May 12, Rogers Youth Center, 315 West Olive Street, Rogers. 8 AM to 4 PM. Doors open 6 AM for setup only. Commercial exhibitors and flea market tables/space $2.00. Free admission. Programs and activities scheduled. Parking on premises. Talk in on 16/76 and 52 simplex. For more information: Roy Milliren, AF5W, 2014 South 16th Street, Rogers, AR 72756.

CALIFORNIA: *The West Coast VHF/UHF Conference* will be held May 4, 5 and 6 at the Paso Robles Inn, Paso Robles. For further info contact: K6HXW, Mike Goshay, PO Box 493, Arroyo Grande, CA 93420.

CONNECTICUT: The seventh annual Pioneer Valley Radio Association (P.V.R.A.) Flea Market, Sunday, April 29, 10 AM to 4 PM at Penney High School, Forbes Street, East Hartford. Talk in on 19/79. For information and reservations contact: Jon Patz, KA1FYL, 34 Whiting Lane, West Hartford, CT 06119 or call (203) 232-8772, evenings.

GEORGIA: The Atlanta Hamfestival 1984, sponsored by the Atlanta Radio Club, June 16 and 17, at the Atlanta Civic Center. 70,000 square feet of air-conditioned exhibitor space and over 800 outdoor flea market spaces will be available. Flea Market $12.50 per space in advance; $15.00 at the gate for both days. Hamfest registration $5.00 in advance, $6.00 at the door. To be pre-registered for the Flea Market or Hamfest, we must receive your application and check by June 8. Pre-registration applications received after June 8 will be returned. Hours 8 AM to 5 PM on Saturday, 8 AM to 2:30 PM on Sunday. Talk in on 3.97 MHz, 146.22/82 and 146.94 simplex. For pre-registration or other information write Atlanta Radio Club, PO Box 77171, Atlanta, GA 30357.

ILLINOIS: M.A.R.K. Hamfest, sponsored by the Moultrie Ama-

teur Radio Klub, May 6, 4-H Fairgrounds, Cadwell Road, Sullivan. Covered facilities, free swappers row, lunch served. Talk in on 14-.655-.055 and 146.520. For more information: WA9WOB (217) 268-3139 evenings.

ILLINOIS: The Chicago Amateur Radio Club's evening Mini-Hamfest, Wednesday, May 2, 6 to 10 PM, Edgebrook Golf Course Field House, 6100 N. Central between Elston and Devon. Admission $1.00. Card table spaces $3.00. Talk in on 146.52 MHz. Refreshments. For info/tickets/space reservations SASE to C.A.R.C., 5631 W. Irving Park Rd., Chicago, IL 60634 or phone (312) 545-3622.

ILLINOIS: A Computer and Electronics Swapmeet, sponsored by Micro International, computer equipment distributors, will be held Sunday, April 15, 10 AM to 5 PM at the American Legion Hall, Butterfield and Spring Roads in Elmhurst. Admission $2.00 advance, $3.00 at the door. For more information contact Micro International, Dept. M, PO Box 774, Highland Park, IL 60035. (312) 530-1552.

ILLINOIS: The Centralia Wireless Association's annual Hamfest, Sunday, May 6, Kaskaskia College Gym, 3 miles northwest of Centralia. No charge for flea market and exhibit space; doors open 7 AM for set up. Free admission and free parking. Refreshments available. Talk in on 147.27/.87 and 146.52. For further information: Bud King, WB9QEG (618) 532-6606, Lou Hodges, W9IL (618) 533-4724 or write CWA, Inc., PO Box 1166, Centralia, IL 62801.

ILLINOIS: The annual Kankakee Hamfest, May 6, Kankakee County Fairgrounds. Gates open 8 AM. Large indoor/outdoor flea market. Refreshments available. FCC booth. Shuttle service available from nearby Greater Kankakee Airport. Advance tickets $2.50; $3.00 at the gate. Talk in on 146.34/.94. For further information call Don Kerouac before 5 PM at (815) 937-2750 or write KARS Hamfest, 1377 Circle Dr. N.W., Kankakee, IL 60901.

INDIANA: The Annual Evansville TARS Hamfest, May 20, all indoors at the Vanderburgh County 4-H Fairgrounds. Open 6 AM CDT. Admission $3.00. Indoor tables $7.50. Outdoor flea market $3.00. Talk in on 147.75/.15 and 146.19/.79. For table reservations and information contact Mike Anderson, KA9LQM, PO Box 3284, Evansville, IN 47732.

MINNESOTA: The 7th annual Rochester Area Hamfest, Saturday, April 7, John Adams Junior High School, 2535 N.W. 31 Street, Rochester. Sponsored by the Rochester ARC and the Rochester Repeater Society. Doors open at 8:30 AM. Large indoor flea market, refreshments and plenty of free parking. Talk in on 146.22/82 MHz. For further information: RARC c/o WB0YEE, 2253 Nordic Ct. N.W., Rochester, MN 55901.

MISSISSIPPI: ARRL Mississippi State Convention/Capital City Hamfest, April 14 and 15, Communications Workers of America Building, Jackson. Dealer exhibits, indoor flea market, concessions, free parking, forums. Free admission. Flea market tables $5.00. Talk in 146.16/76 MHz. Contact Carol Kemp, NA5Y, 3581 Beaumont Drive, Pearl, MS 39208. (601) 939-7612.

MISSOURI STATE ARRL Convention, Kansas City, April 7-8, 1984. For information write or call PHD Amateur Radio Association, PO Box 11, Liberty, MO 64068-0011. (816) 781-7313.

NEW ENGLAND: The Hosstraders Spring Tailgate Swapfest, Saturday, May 12, sunrise to sunset at Deerfield, NH. Fairgrounds. Admission $2 includes tailgaters and commercial. Friday night camping for self-contained rigs at nominal fee. None admitted before 4 PM Friday. No reserved spaces. Profits benefit Boston Burns Unit of Shriners' Hospital. Last year's donation over $4700.00. For map to northeast's biggest Ham Flea Market SASE to Norm, WA1IVB, RFD Box 57, West Baldwin, ME 04091 or Joe, K1ROG, Star Route Box 56, Bucksport, ME 04416 or Bob, W1GWU, Walton Road, Seabrook, NH 03874.

NEW HAMPSHIRE: Springfest '84, the 4th annual Hamfest-Flea Market sponsored by the Great Bay Radio Association, Saturday, April 7, 9 AM to 3 PM, Rochester V.F.W. Post 1772 Hall, Pickering Rd., Rochester (Gonic). Food and refreshments available. Free parking. Admission $1.00. Talk in on 147.57. For table reservations and information write: Great Bay Radio Assn., PO Box 911, Dover, NH 03820.

NEW JERSEY: TCRA Hamfest Tri County Radio Association, rain or shine, Sunday, May 13, Passaic Valley Community Center off Valley Road, Stirling, NJ. 9 AM to 4 PM, indoors, refreshments, rest rooms, free parking. Tables $6, registration $2.50. Table reservations call or write Dick Franklin, W2EUF (201) 232-5955 or 270-3193, PO Box 182, Westfield, NJ 07090.

NEW YORK: Suffolk County Radio Club Indoor/Outdoor Flea Market, Sunday, May 6, 8 AM to 3 PM, Republic Lodge No 1987, 585 Broadhollow Road (Rt. 110), Melville. Refreshments, free parking. General admission $2.00. Wives and children under 12 free. Indoor sellers' tables $7.00. Outdoor space $5.00 (includes one admission). Talk in on 144.61/145.21 and 146.52. For additional information contact: Richard Tygar, AC2P, (516) 643-5956 evenings.

NEW YORK: The 25th annual Southern Tier Amateur Radio Club's Hamfest, Saturday, May 5, Treadway Inn, Owego. Flea market opens 8 AM. Displays, tech and non-tech talks, refreshments. There will be a dinner at 6:30 PM by advance tickets only. Talk in on 22/82, 16/76 or 146.52 simplex. For further information SASE to KF2X, C. England, RD #1, Box 144, Vestal, NY 13850.

OHIO: The Dayton-Cincinnati Chapter of the Quarter Century Wireless Association will hold its annual banquet during the Dayton Hamvention, Friday, April 27, Neil's Heritage House Restaurant, 2189 S. Dixie Drive, Dayton. C.O.D. bar opens at 6:30 PM, dinner 7:30 PM. Dr. Jerrold Petrofsky, developer of computerized equipment which enables paraplegics to walk, will give an illustrated presentation. Tickets $12.50. Contact Doug Horner, W8PH for details or call (513) 859-3210.

OHIO: The 15th annual B*A*S*H will be held Friday evening at the Dayton Hamvention, April 27, Convention Center, Main and Fifth Streets. Parking in adjacent City Garage. Free admission. Sandwiches, snacks and C.O.D. bar available. Live entertainment. Two exciting top awards and many others. For further information: Miami Valley F.M. Association, PO Box 263, Dayton, OH 45401.

SOUTH CAROLINA: The Blue Ridge Amateur Radio Society's Greenville Hamfest, Saturday, May 5 and Sunday, May 6, American Legion Fairgrounds, White Horse Road 1/2 mile north of I-85 in Greenville. Admission $4.00. Advance tickets $3.00. Talk in on 146.01/61. For information: Phil Mullins, WD4KTG, Chairman, PO Box 99, Simpsonville, SC 29681. For advance sales: Sue Chism, N4ENX, Rt. 6, 203 Lanewood Drive, Greenville, SC 29607.

TEXAS: The San Antonio Area Radio Club's FIRST annual Swapfest and Bar-B-Q, April 7 at Comanche Park, 7 AM to 5 PM. Talk in on 147.36 MHz. For details: Melvin Anderson, 8932 Saddle Trail, San Antonio, Texas 78255.

WISCONSIN: The Ozaukee Radio Club's 6th annual Swapfest, Saturday, May 5, 8 AM to 1 PM, Circle B Recreation Center, Highway 60, Cedarburg. Admission $2.00 advance, $3.00 at the door. 6' tables $2.00; 8' tables $3.00. Food and refreshments. Sellers admitted at 7 AM for set up. For tickets, tables, maps or more information send business SASE to 1984 Ozaukee Radio Club Swapfest, PO Box 13, Port Washington, WI 53074.

OPERATING EVENTS

"Things to do..."

MAY 4: The Gloucester County Amateur Radio Club will operate W2MMD on May 4, 1700Z to May 5, 1700Z to commemorate the Club's 25th anniversary. Phone operation in lower portion of General Class bands 80-10 and CW in Novice bands. Commemorative certificate by QSL to G.C.A.R.C., PO Box 370, Pitman, NJ 08071.

MAY 12: The Greater Fairfield (CT) ARA will operate WB1COO from 1300 UTC to 2200 UTC during the annual Dogwood Festival. Certificate for SASE. Frequencies 3.975, 7.235, 14.330, 21420.

APRIL 5-8: Dordtse Electronica Club members will be active from Luxembourg (LX) on all bands including WARC and 160 from 1800 GMT til 2400 GMT. QST is Bech. They will all sign /LX. QSL via PO Box 523, Dordrecht, the Netherlands or via bureau.

APRIL 21-22: QRP Amateur Radio Club International QSO Party, Saturday, April 21, 1200 UTC to 2400 UTC, Sunday, April 22. Exchange: Members give RS(T), state/province/country and WRP ARCI membership number. Non-members give RS(T), state/province/country and power output. QSO points (total all bands) times total number of states/provinces/countries (may be worked on more than one band) times power multiplier times bonus multiplier (if any) equals claimed score. Send large SASE or IRCs to contest chairman for scoring summary sheet in advance of contest. Send full log data plus separate worksheet showing details and time off air. No logs returned. For results and scores send large SASE or ICSs. Logs must be received by May 21, 1984. QRP ARCI Contest Chairman, Eugene C. Smith, Jr., KA5NLY, #16 Fairmont Drive, Little Rock, AR 72204.

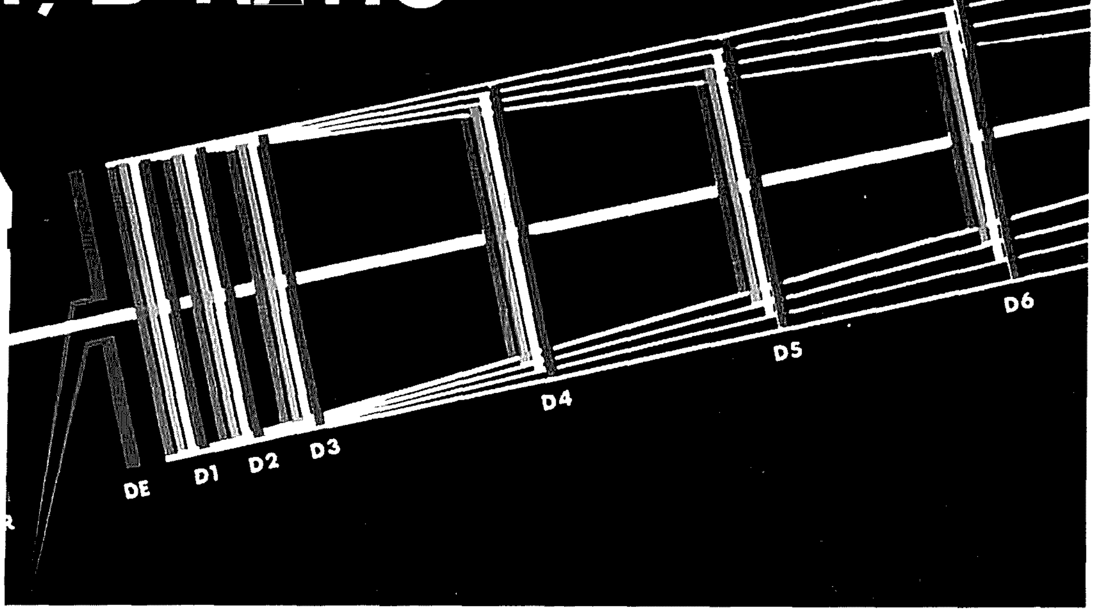
MAY 5-6: Cape Canaveral — the Florida Solar Energy Center of the State University system of Florida and the Indian River ARC will celebrate SUN DAY, 1500Z to 2200Z. 5,880 photovoltaic 4" diameter solar cells will provide heating, cooling, cooking and Amateur Radio operations during this event. W4NLX/4 will operate on SSB 7.240, 14.240, 21.370 and 28.518. CW 7.040, 14.040, 21.040 and 28.003. FM 146.28/88. A beacon will be on 1296.05 MHz. For all SWL and Amateur Radio operators the quarterly high tech newsletter "The Solar Collector" is available free on request. Also a multicolor certificate. Send business SASE to FSEC, 300 State Road 401, Cape Canaveral, FL 32920.

MAY 1984 / $2.50

ANNUAL ANTENNA ISSUE

ham radio magazine

YAGI ELEMENT TAPER AFFECTS GAIN & F/B RATIO



also: remote-controlled 40/80/160-meter vertical • end-fed multiband 8JK • computerized antenna matching • high-performance dipole • branch-line hybrids • simple wire plow • and the conclusion of K2BT's series on vertical phased arrays

communications technology

ham radio
magazine

MAY 1984

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contents

REFLECTIONS

One year ago . . .

It's hard to believe a year has passed since we put together the 1983 annual antenna issue. That was my first May special as Editor-in-Chief (though I had joined *ham radio* in the fall of the previous year). Gracing the pages of that issue was the start of a series of articles on vertical phased arrays that promised to examine in detail every important aspect of that subject from design through construction. Most appropriately, the final article in that series appears in this special antenna issue. I would not for a moment suggest that it is either easy reading or the answer to everyone's antenna requirements, but as the author has repeatedly stated, the series takes the mystery out of antenna development and squarely places that topic into the realm of science. Though the author has primarily concentrated on the 75-meter band (where his interests lie — and considering the present sunspot cycle trend, it's not a bad idea), the concept he presents is well suited for use on other bands. It's a series I heartily recommend reading and rereading.

While still on the subject of verticals, how would you like to be able to remotely control a fine performing antenna on your three favorite low bands: 160, 80, and 40? Robert Leo, W7LR shows you how with his 70-foot vertical that uses inexpensive irrigation pipe and a five position switching control head and RF deck. It's simple and effective, and with a flick of a switch, gives you *full* coverage of the three bands without any retuning . . . a very nice addition, I might add, to any installation that uses a no-tune synthesized transceiver.

From the United Kingdom comes an ingenious and quite effective design for an antenna and tuner that enables operation on the three highest bands (20, 15, and 10). R.C. Marshall, G3SBA, shows how an old standby antenna, an "8JK," can be end-fed to provide performance comparable to that of the standard center-fed version. The more remarkable aspect of his development is a sophisticated antenna control unit that enables the operator to step through each of the bands in narrow segments using a BCD switching circuit. The original motivation for this elaborate design was the author's need for an unobtrusive, high performance three-band antenna installation. Food for thought for those with small lots and big ambitions (DX, that is).

On a different angle, *ham radio* received an innocent enough sounding call several months ago from a gentleman who said that he had carefully read Lawson's series on Yagis for the HF bands and would be interested in providing a somewhat similar presentation for the VHF/UHF bands. From subsequent conversations with Stan Jaffin, WB3BGU, it soon became obvious that not only had he thoroughly understood Lawson's series, but that he'd also investigated the works of several prominent VHF/UHF designers including Kmosko and Johnson, Greenblum, Tilton, Viezbicke and Knadle. With the aid of a computer (a mainframe, I believe), he generated a series of tables and associated patterns which examined in fine detail the effect of iterating Yagi parameters for the enhancement of gain and front-to-back ratio. His first article, appearing in this issue, is concerned with the 2-meter band with emphasis on the weak signal segment. Very methodically various director tapering schemes are shown and logical conclusions drawn. The series will continue over the next several months, with the same care given to the examination of Yagis for the 220 and 432 MHz bands. Like Lawson's series, Stan's is not meant for casual reading, but rather for use by experimenters who wish to build a strong foundation in their understanding of Yagi antenna design.

While in the VHF/UHF World; Joe Reisert, W1JR, one of our featured columnists, has done it again. The no-nonsense delivery of information he's famous for has carried into this special antenna issue. In his column on page 110, he examines the "slightly" important area of performance parameters. Gain, beamwidth, front-to-back ratio, sidelobes and VSWR are defined and their interrelationships explored. He doesn't stop there, but continues to discuss other areas of considerable importance to the antenna designer such as feed systems, wind load, structural strength and preventive maintenance. In the second part of his article — of special interest to those who attend antenna measuring contests — Joe shows how one can actually determine gain and come pretty close with an estimate once the major side lobe levels are known. In reading the article perhaps you'll recognize several of the common mistakes that can turn a VHFer's hair a lighter shade of white.

Sometimes it appears that very few new antenna ideas are generated, and that recent developments are old concepts introduced to meet new needs. A case in point is what's known as a "stretch" or phase compensated dipole. The concept, first introduced in the 40s and possibly a decade earlier, illustrates a technique whereby the addition of series capacitors, periodically located throughout an antenna's legs, create a uniform current distribution over the entire length of the radiator. The particular version explored by David Atkins, W6VX, sports a low angle of radiation, high efficiency and wideband performance.

This annual antenna issue continues with techniques using just a GDO, noise or RF bridge — or more elaborately, a microcomputer — to match simple antennas.

For those intrigued with verticals, a large ground radial system is a must and W7IV shows you how it can be done with the use of a home-built wire plow.

Naturally pleased with this month's useful editorial content, we also take considerable satisfaction in setting a new *ham radio* record: because of the outstanding support of you, our readers, and our advertisers, this issue sets an all-time record for advertising space sales. To all involved, a hearty THANK YOU.

Remember, I always have room for just one more manuscript. If you have an idea for an article you'd like to write, send it in or stop by the *ham radio* booth at the Dayton Hamvention, April 27-29. Let's talk about it.

Rich Rosen, K2RR/1
Editor-in-Chief

presstop de W9JUV

TOP ARRL OFFICES WERE SHAKEN UP IN A STRONGLY CONTESTED ELECTION March 26. Elected to League presidency was First Vice President Larry Price, W4RA, while Great Lakes Division Director Len Nathanson, W8RC, was named First Vice President. Gar Anderson, KØGA, retained his Vice Presidency, while Southwestern Division Director Jay Holladay, W6EJJ, was also elected a Vice President. Out entirely is Carl Smith, WØBWJ, who had moved up to President from First Vice President after the death of Vic Clark, W4KFC, in November.

Just What This Means To The League's Future is hard to gauge as yet, though some knowledgeable League observers believe it's just the beginning of a period of internal strife and major policy shifts. It's no secret that many of the staff have been very unhappy over recent changes in work schedules, and the new administration is not believed to be very sympathetic to their complaints.

AN UNCOORDINATED REPEATER HAS BEEN SHUT DOWN by the FCC in southern California. The action came after the operator of a coordinated repeater (to which the uncoordinated repeater had caused interference) and the area coordinator, TASMA (the Two Meter Area Spectrum Management Association), filed complaints with the FCC. Cited in the FCC's action was a policy letter written last April by the FCC's Jim McKinney to WA2DHF of the Tri-State Repeater Group, in which he stated that in cases of interference the FCC would support coordinated repeaters and the organizations that provided their coordination.

IT'S ALMOST A HORSE RACE TO SEE WHICH VEC WILL ADMINISTER the first Amateur exams under the new Volunteer Exam Program. DARA plans to give exams at the Dayton Hamvention the last weekend in April, the Anchorage Amateur Radio Club is believed to be on about the same timetable, and the DeVry ARS plans its first exam session by May 8 at DeVry's Chicago campus. The proposal of a sixth VEC, "VEC Region Four, Inc.," has also been accepted for the fourth call area, but a fifth district group's proposal has been returned for revision.

The FCC's Proposal To Permit Reimbursement Of VEC Expenses was released March 9, with a Comment due date of April 16 and Reply Comments due May 1. In it the Commission proposed letting either the VEC or VE collect the fee, with the VEC setting up procedures by which VEs could also be reimbursed for their expenses. Despite the unusually short comment period on this NPRM, Commission procedures make it unlikely that VECs can actually begin collecting fees before late summer. Since the ARRL directors have taken the firm position that they won't make their VEC proposal to the FCC until fees are in place, other VEC programs will be up and running in much of the country before the League can even start its own. Even the work the League has already done on the volunteer program isn't being utilized. When one of the newly appointed VECs approached the League recently with the suggestion that the VECs already in place and the League share their efforts on the exam question and answer pool, in order to assure uniform exams throughout the country, that VEC was told "the League has invested a lot of time and money in that effort and doesn't feel it can share it!" A further unknown factor is the new League administration, which could make radical changes in the direction the League is taking in the VEC program.

A Volunteer Amateur Exam Program In Those Parts Of The Country not yet represented by a VEC will become a must before much longer, with or without the ARRL. The FCC's intention is to get out of Amateur license exams entirely before the end of 1984, with the exception of possible special review exams for Amateurs whose licenses are suspect.

THE CHICAGO AREA FCC-AMATEUR 2-METER INTERFERENCE COMMITTEE held its second meeting March 14, drawing repeater representatives from all over northern Illinois and southeast Wisconsin. DFing techniques and equipment, both Amateur and FCC, were discussed at some length, and the Commission representatives distributed forms they'd like used for documentation and reporting of jamming problems.

This Program Is Being Watched By Washington As A Possible Prototype for the long-awaited program to incorporate Amateur volunteers in the FCC's enforcement effort. At present the Chicago Field Office wants to know about on-going interference problems, but on an "organized" basis through designated representatives of each repeater.

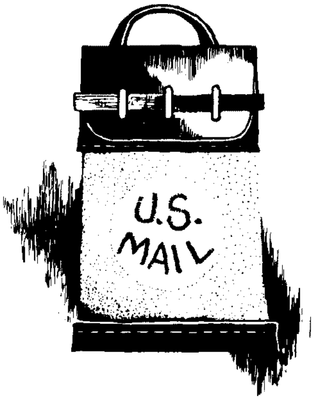
The Group Voted Unanimously To Continue The Program, and expressed very positive sentiments toward the Chicago Field Office people and their willingness to work with and for the area Amateur community. The next meeting will probably be held in June.

FULL PRIVILEGES ON THE 1900-2000 kHz PORTION OF 160 METERS were restored to U.S. Amateurs in a March 22 FCC action, which noted that Canadian Loran-A operation on 160 has ended.

W2NSD's PROPOSAL TO REQUIRE CW RETESTING OF ALL AMATEURS has been rejected by the FCC as being "without merit" and "not presenting any new or novel issues...."

A PROPOSAL TO GIVE NOVICES ALL-MODE PRIVILEGES ON 220 MHZ is being prepared for FCC submission in mid-May by WA2MCT/5 and WD5DON. Rationale for the proposal, which would limit Novices to 233.40-233.75 MHz, is to improve usage of the band while allowing Novices to take part in public service activities and gain "hands-on" experience with other modes.

AMATEUR STATIONS WILL BE PERMITTED IN THE OLYMPIC VILLAGES in Los Angeles after all, it now appears. After drawn-out, on-and-off negotiations due largely to concerns about security, the Olympic Committee has agreed to permit three stations in the competitor's compounds. A limited number of volunteer operators, all subject to detailed security clearances, will be permitted to operate the stations and provide a communications link back home for those Olympic competitors whose countries permit it.



comments

smoke signals

Dear HR:

In response to Fred Norvick's comment on heat sinks (Comments, October, 1983), the use of a capacitor in the power line feeding a heat sink cooling fan might cause the fan to increase speed to the point that it might start sending "smoke signals." If the added capacitance and the inductance (of the fan) are (series) resonant at 60 cycles, only the ohmic resistance remains in the circuit, causing the fan to draw more current than it is designed for. You will have to experiment with different capacitor values to find the one that will make the fan rotate more slowly. I increase fan speed in my 2-meter linear (4X250A) during the SSB transmit cycle and let it "coast" during receive periods.

Rudolf Frank, DJ4BZ
Burgkirchen, Germany

AM update

Dear HR:

In Bill Orr's comments on "ancient modulation" in the February issue of *ham radio* (page 65), Bill correctly points out how Amateur Radio voice quality has deteriorated since the advent of SSB. Today, even on VHF-FM, a medium fully capable of extremely good audio quality, voices typically sound more like they are coming from tin can telephones than from sophisticated communications equipment. Many of the most expensive HF and VHF transceivers, loaded with all the latest "bells and whistles," come equipped with cheap CB-style hand mikes and tiny speakers. The prevailing philosophy seems to have been that in Amateur Radio, voice quality does not matter. Certainly we can do better than that.

I must take issue with one statement in Bill's article, however. Users of AM have not "gradually retreated to obscure regions of 160 and 10 meters." AM activity can be heard daily on most of the HF bands. AM'ers, now a minority, tend to operate within certain portions of the bands, much in the same manner as RTTY and SSTV operators. The most commonly used AM frequencies are 1880-1900, 1985, 3860-3890, 7160, 7285-95, 14286, and 29,000-29,200 kHz. If anything, there has been a renewed interest in Amateur AM in recent years. It is interesting to note in looking through the ads that AM is being increasingly included on the newer "all-mode" transceivers. Moreover, some of the latest rigs are capable of very decent voice quality on AM and SSB. The Amateur equipment market is extremely cost competitive, yet the inclusion of AM on a transceiver must add considerably to its selling price. The major Amateur equipment manufacturers are large companies with worldwide markets. It is inconceivable that marketing research data would not be a factor in the design of their products. These manufacturers simply would not drive up the cost of their transceivers by including the AM mode if they were not convinced that there is a substantial demand for Amateur equipment with AM capability.

Most AM'ers are particularly interested in voice quality; some of the AM signals heard on the ham bands would put many broadcast stations to shame. There has already been some Amateur experimentation with pulse duration modulation, and AM'ers routinely use advanced techniques such as equalizers, delay/reverb (not CB-style echo boxes), and other processing techniques. Many AM operators use older equipment they have refurbished themselves, and there are even some home-built stations on the air! Modern AM techniques have not completely bypassed Amateur Radio. The "Amateurs" in our ranks who deride those of us who operate and experiment with amplitude modulation usually turn out to be individuals who take pride in their ignorance.

Donald Chester, K4KYV
Woodlawn, Tennessee

volunteer examiners: keep standards high

Dear HR:

I am concerned about the honesty of the Volunteer Examination Program. Right now we have many licensed "Amateurs" who could not pass an honest General Class examination. I have suggested to the ARRL that all volunteer examiners be Extra Class with at least 10 years of experience, and furthermore, that the volunteer examiners be required to affirm and certify that they meet the requirements for Extra Class in all respects, including code and theory, and that improper conduct in the administration of the examination would lead to license revocation. The ARRL has not yet responded to these suggestions.

The government recognizes that the skilled Amateur is a valuable asset. During the war all of the Electronic Field Engineers at Raytheon were Amateurs. (Clark Rodimon from ARRL Headquarters headed up this group.) The four senior people in the Bureau of Ships Radar Design Branch were Amateurs. At that point you knew what to expect of an Advanced or General Class Amateur, but I doubt if the same standards are applicable today.

I.L. McNally, K6WX
Sun City, California

SHORT CIRCUIT HOTLINE

Building a current *ham radio* project? Call the Short Circuit Hotline any time between 9 AM and Noon, or 1 to 3 PM — Eastern time — *before* you begin construction. We'll let you know of any changes or corrections that should be made to the article describing your project.

(See "Publisher's Log," April, 1984, page 6, for details.)

603-878-1441

applied Yagi antenna design
part 1: a 2-meter classic revisited

Computer model analyzes, updates Kmosko-Johnson designs

The Yagi antenna model developed by the late James Lawson, W2PV, provided a rigorous method for exploring Yagi antenna design and performance.[1] Because his primary interest was the HF region, Lawson's nine-part series[2] emphasized the two- through six-element Yagis that, because of boom length, are useful mainly below 30 MHz; the ability of his model to measure the performance of the multi-element and multi-wavelength Yagis found in VHF and UHF applications, however, was also discussed.[3]

This series adapts Lawson's model for use with specific VHF and UHF antenna designs. A computer model is used to optimize forward gain and front-to-back (F/B) ratios for the weak signal area of given VHF/UHF bands. Each resulting antenna design can be polarized either horizontally or vertically, or stacked for enhancement of specific performance parameters.

basic technical parameters

Several technical assumptions underlie the analyses in this series. First, all Yagi antenna designs are based on non-conductive booms. Methods for conversion of element lengths to round or square conductive booms are readily available.[4] Second, a non-reactive driven element is used; this means a self-impedance value of 73 + j0. Third, antenna feeding methods are left to the user's discretion; any number of proven feeding methods are available for VHF and UHF Yagis.[5] Finally, wherever it is possible, common intervals such as 0.125 or 0.0625 inches are used for Yagi antenna elements or iterations. Few Radio Amateurs can readily measure and cut antenna elements to smaller tolerances, and the use of computer iteration assures finding optimized lengths at these intervals. However, as stated above, the driven element is an exception and will be stated to the nearest 0.000001 inch.

The purpose of this series is to provide analyses of Yagi antenna designs for application above 50 MHz. There is no real limit to the number of iterations that can be run against a given antenna design. However, the practical side of any design effort requires that some sort of sampling be made, particularly in terms of the analyses that are performed and the select few that are reported. This is the rationale for limiting this first article to six variations of a basic design, with each variation presented in terms of gain and F/B optimization.

the classic Kmosko-Johnson design

One of the best known designs for a 144 MHz antenna was published 28 years ago.[6] Combining variable director spacing, tapered elements, and a 3.44 wavelength boom, this design represents the results of a long empirical process on the part of James Kmosko, W2NLY, and Herbert Johnson, W6QKI, the designers. Another version of this same antenna, but with different reflector length and spacing as well as a different element mounting method, was published later.[7] As the original design is more widely known, it is the one selected for computer analysis.

Table 1 contains the original Kmosko-Johnson antenna dimensions and wavelength values for 144.0 MHz. The element diameter is 0.09375 inches, and each element is spotwelded across the diagonal of a 0.75 inch square metal plate. The plate is screwed to the top of a 1.25 inch diameter aluminum boom.

Conversion of this antenna to its non-conducting boom equivalent presented several problems. While information for preliminary estimates was available,[8] it was decided to sidestep the issue via the use of the brute force computer iteration method.

As the design frequency was also to be shifted to 144.5 MHz, initial iterations of reflector and director lengths were made across a wide range of values. The element diameter remained as in the original design; two samples of the intermediate results are presented in table 2 and table 3.* Because the designers

*In reading these and the tables that follow, it should be noted that only the length of the first director is specified. Each table's title contains the tapering scheme to apply to subsequent directors.

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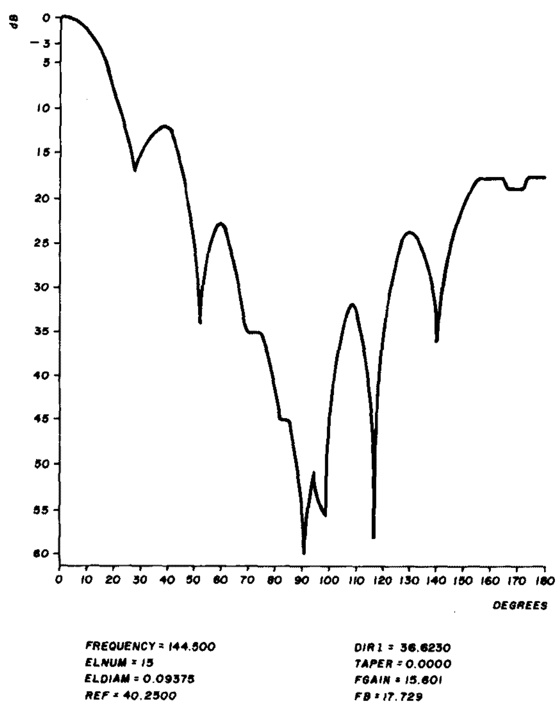

fig. 1. Plot of the zero taper, gain optimized antenna.

table 1. Original Kmosko-Johnson antenna design for 144.0 MHz.

element		length (inches)	element spacing (λ)	cumulative length (λ)*
1	reflector	41.500	0.0000	0.0000
2	driven	39.500	0.2440	0.2440
3	director 1	37.750	0.0854	0.3294
4	director 2	37.625	0.0915	0.4209
5	director 3	37.500	0.0915	0.5124
6	director 4	37.375	0.1952	0.7076
7	director 5	37.250	0.3904	1.0980
8	director 6	37.125	0.3904	1.4884
9	director 7	37.000	0.3904	1.8788
10	director 8	36.875	0.3904	2.2692
11	director 9	36.750	0.3904	2.6596
12	director 10	36.625	0.3904	3.0500
13	director 11	36.500	0.3904	3.4404

*cumulative length (λ) of antenna.

asserted that their non-tapered antenna was in fact the optimal gain antenna, locating the parasitic element lengths for maximum gain at the new design frequency would result in a comparable antenna.

In terms of the actual antenna undergoing computer iteration, the designers specified that to move the design frequency within the 2-meter band, only the element lengths had to be changed. The spacing between the elements for frequencies between 144.0 and 148.0 MHz remains the same as at 144.0 MHz. Therefore, the experimental 144.5 MHz antenna was used as a standard for comparison, as shown in **table 4**. The apparent changes in element spacings are due to slight shortening in wavelength at the new design frequency. The lengths of the parasitic elements are supplied with each iteration.

Frequency response parameters are provided for each of 12 optimized antennas. Since 144.5 MHz is the design frequency and the designers specifically showed how performance was made to drop dramatically above 146.0 MHz, a range of 142.5 to 146.5 MHz is used. Each frequency performance table contains nine data points, with 144.5 MHz as the frequency center.

computer-designed Kmosko-Johnson antennas

Performance measurements were supplied for tapers of 0.000, 0.125, and 0.25 inch. The last was a special taper wherein the first three directors used a 0.125 inch taper and the remaining directors tapered at 0.25 inch.

In order to analyze the Kmosko-Johnson design more thoroughly, and to illustrate the ease and utility of computer-based Yagi antenna design, three additional tapers (0.0625, 0.1875, and 0.25 inch) are included. Here, 0.25 is a linear taper applied equally to directors two through eleven.

E-plane (cartesian) plots (**figs. 1-13**) are presented for the Yagis in each tapering procedure and are calculated at 144.5 MHz. A computer line printer served as the output device, with the attendant limitations of the standard 11 × 14 7/8-inch page size. The dB range is 0 to 60, with calculated values in excess of 60 shown as 60 dB. For some antennas this may indicate a trough instead of a null, but little accuracy is lost as nulls of this magnitude are rarely achieved in practice. The degree range is 0 to 180, with every second degree (0,2,4,6, and so on) being plotted. Because computer printers are discrete output devices, only full integer dB values can be plotted, meaning that 10.75 dB at 100 degrees is printed with an "E" (for E-plane) at 11 dB at 100 degrees. A skilled artist (WA9MXB) has drawn curves in place of the discrete E's; these are the fine plots that appear as **figs. 1** through **13**. The plots in the rest of this series are generated in a similar manner.

General observations are presented after the discussions of the major attributes of each of the six tapers. As has often been the case with antennas, there are no absolutes. Often the "best" antenna is more a function of the station operator than of any of the antenna's electrical or mechanical attributes.

taper = 0.000

The designers specified that a zero taper resulted in an optimized forward gain and a comparatively poor F/B. **Table 5** depicts the iteration producing the zero taper's optimum gain of 15.601 dBi (**fig. 1**), and **table 6** depicts the iteration producing the F/B optimization of 32.918 dB (**fig. 2**). The differences in reflector and director length indicate that these are in fact two

table 2. Iteration of the Kmosko-Johnson antenna with a reflector length of 39.75 inches at 144.5 MHz.

director 1 (inches)	gain (dBi)	F/B (dB)
36.500	15.199	11.426
36.625	15.277	11.559
36.750	15.351	11.796
36.875	15.421	12.174
37.000	15.486	12.750
37.125	15.542	13.619
37.250	15.583	14.952
37.375	15.593	17.112
37.500	15.546	21.106
37.625	15.385	32.100
37.750	14.998	23.112

table 3. Iteration of the Kmosko-Johnson antenna with a reflector length of 41.375 inches at 144.5 MHz.

director 1 (inches)	gain (dBi)	F/B (dB)
36.500	14.968	12.975
36.625	15.075	13.204
36.750	15.182	13.545
36.875	15.288	14.038
37.000	15.388	14.745
37.125	15.476	15.766
37.250	15.538	17.285
37.375	15.577	18.953
37.500	15.492	23.907
37.625	15.302	29.637
37.750	14.903	22.009

table 4. Baselined antenna at 144.5 MHz with fixed parasitic element spacings and parasitic element lengths supplied during iteration.

	element	length (inches)	element spacing (λ)	cumulative length (λ)
1	reflector	—	0.0000	0.0000
2	driven	39.320721	0.2449	0.2449
3	director 1	—	0.0857	0.3306
4	director 2	—	0.0918	0.4224
5	director 3	—	0.0918	0.5142
6	director 4	—	0.1959	0.7101
7	director 5	—	0.3918	1.1019
8	director 6	—	0.3918	1.4937
9	director 7	—	0.3918	1.8855
10	director 8	—	0.3918	2.2773
11	director 9	—	0.3918	2.6691
12	director 10	—	0.3918	3.0609
13	director 11	—	0.3918	3.4527

fig. 2. Plot of the zero taper, F/B optimized antenna.

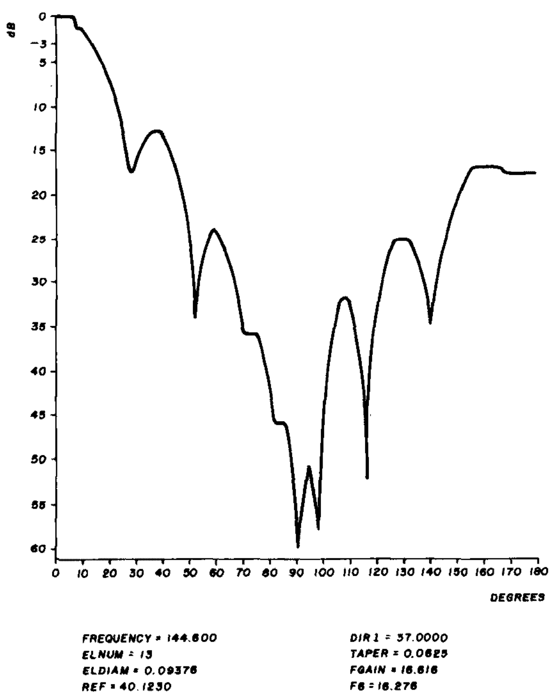

fig. 3. Plot of the 0.0625 taper, gain optimized antenna.

table 5. Maximized gain iteration for taper of 0.000 with a reflector length of 40.250 inches.

director 1 (inches)	gain (dBi)	F/B (dB)
36.500	15.540	15.365
36.625	15.601	17.729
36.750	15.543	19.153
36.875	15.190	15.053
37.000	14.230	9.410
37.125	11.340	6.648
37.250	9.259	−0.370†
37.375	5.347	−4.911
37.500	1.635	−8.585
37.625	0.580	−8.842
37.750	1.319	−7.336

table 6. Maximized F/B iteration for taper of 0.000 with a reflector length of 38.750 inches.

director 1 (inches)	gain (dBi)	F/B (dB)
36.500	13.248	21.148
36.625	15.190	17.977
36.750	15.069	32.918
36.875	14.744	18.295
37.000	13.996	9.756
37.125	12.430	4.130
37.250	9.601	−0.571
37.375	5.489	−4.758
37.500	1.348	−7.580

table 7. Frequency response parameters for gain maximized antenna with a 0.000 taper.

frequency	gain (dBi)	F/B (dB)
142.5	15.259	11.162
143.0	15.344	12.022
143.5	15.440	13.266
144.0	15.537	15.126
144.5	15.601	17.729
145.0	15.524	18.591
145.5	15.082	13.846
146.0	13.920	8.321
146.5	11.736	3.390

table 8. Frequency response parameters for F/B maximized antenna with a 0.000 taper.

frequency	gain (dBi)	F/B (dB)
142.5	14.872	9.832
143.0	14.989	11.304
143.5	14.508	12.832
144.0	15.104	17.813
144.5	15.069	32.918
145.0	14.844	18.294
145.5	14.113	9.726
146.0	12.310	3.985
146.5	9.032	−0.944

† This computer model calculated negative F/B because the rear lobe radiated greater amplitude than the forward lobe.

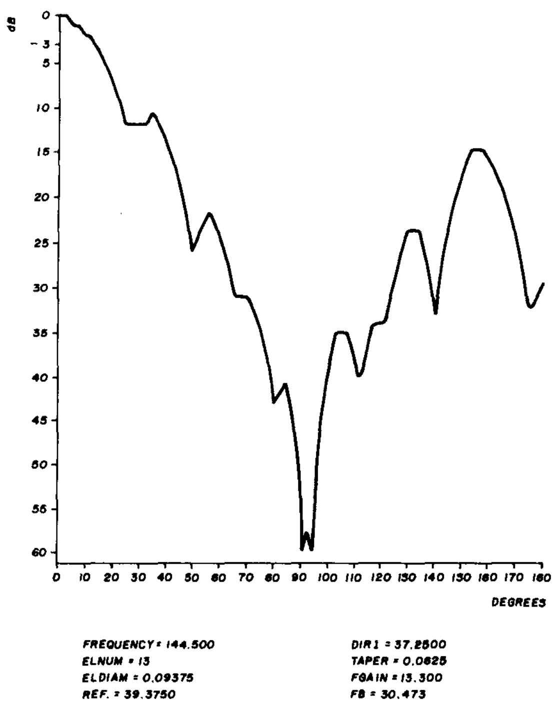

fig. 4. Plot of the 0.0625 taper, F/B optimized antenna.

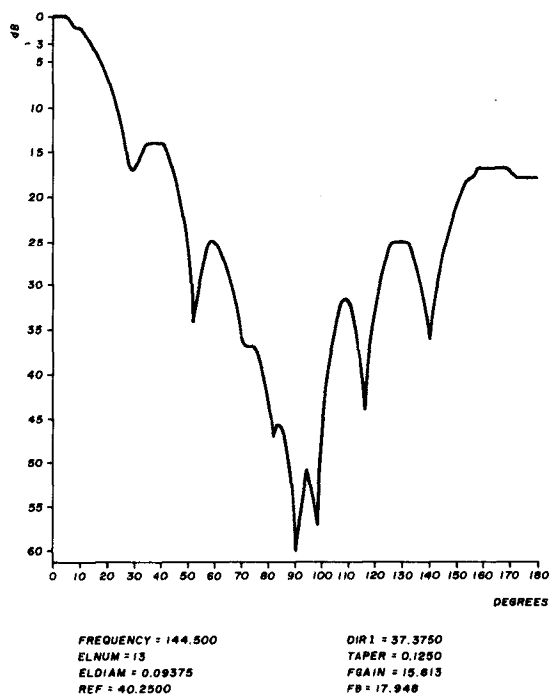

fig. 5. Plot of the 0.125 taper, gain optimized antenna.

fig. 6. Plot of the 0.125 taper informational antenna.

significantly different antennas. **Table 7** presents the optimized gain antenna's computer performance across the 4 MHz bandwidth, and **table 8** does likewise for the F/B optimized antenna. Using this taper, a superior design frequency F/B ratio can apparently be obtained at a sacrifice of only 0.5 dB in gain.

taper = 0.0625

This antenna was not presented by the designers and is totally a product of computer iteration. **Table 9** presents this antenna's optimized gain iteration for 15.618 dBi (**fig. 3**), and **table 10** does likewise for the optimized F/B iteration for 30.473 dB (**fig. 4**). As was the case with the zero taper, these two optimized antennas are significantly different antennas. **Table 11** presents the gain optimized antenna's performance across the 4 MHz bandwidth, and **table 12** does likewise for the F/B optimized antenna. Compared to the zero-taper antenna, this antenna can be optimized to a slightly higher gain or to a higher F/B ratio within the weak signal area part of the band.

taper = 0.125

Of the three antennas they presented, the designers specified this antenna (**fig. 5**) as their best all-around performer. **Table 13** *presents the optimization iteration* that produced the highest gain and F/B calcula-

table 9. Maximum gain iteration for taper of 0.0625 with a reflector length of 40.125 inches.

director 1 (inches)	gain (dBi)	F/B (dB)
36.500	15.345	12.341
36.625	15.454	11.532
36.750	15.510	13.955
36.875	15.581	15.538
37.000	15.618	18.276
37.125	15.573	23.686
37.250	15.344	24.391
37.375	14.742	15.243
37.500	13.474	9.103
37.625	11.192	4.135
37.750	7.587	−0.538

table 10. Maximum F/B iteration for taper of 0.0625 with a reflector length of 39.375 inches.

director 1 (inches)	gain (dBi)	F/B (dB)
36.500	15.330	11.114
36.625	15.399	11.803
36.750	15.460	12.798
36.875	15.510	14.395
37.000	15.533	17.195
37.125	15.491	23.497
37.250	15.300	30.473
37.375	14.778	15.783
37.500	13.579	9.177

table 11. Frequency response parameters for gain maximized antenna with a 0.0625 taper.

frequency	gain (dBi)	F/B (dB)
142.5	15.341	11.401
143.0	15.422	12.240
143.5	15.503	13.435
144.0	15.577	15.248
144.5	15.618	18.276
145.0	15.565	23.786
145.5	15.288	22.055
146.0	14.580	14.019
146.5	13.168	8.283

table 12. Frequency response parameters for F/B maximized antenna with a 0.0625 taper.

frequency	gain (dBi)	F/B (dB)
142.5	15.306	11.839
143.0	15.391	13.622
143.5	15.451	16.556
144.0	15.455	22.877
144.5	15.300	30.473
145.0	14.756	15.518
145.5	13.421	8.826
146.0	10.878	3.697
146.5	6.892	−1.099

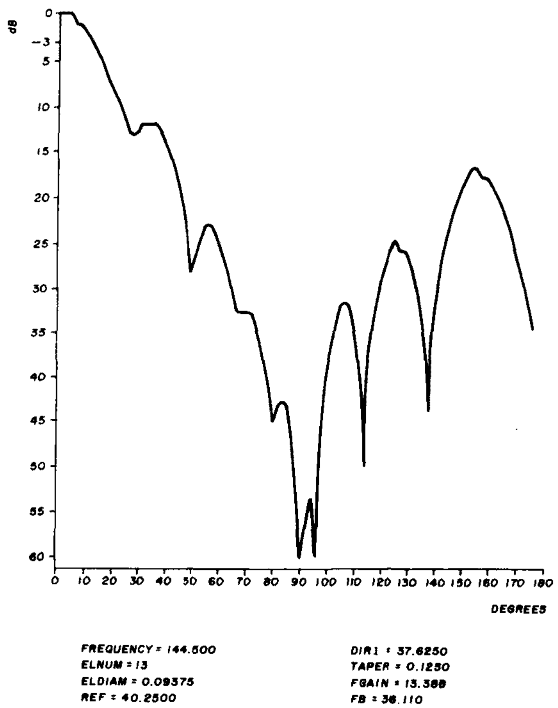

fig. 7. Plot of the 0.125 taper, F/B optimized antenna.

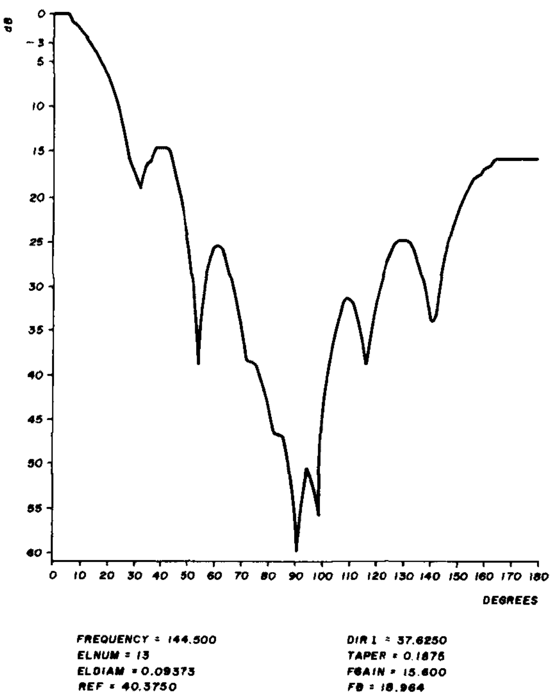

fig. 8. Plot of the 0.1875 taper, gain optimized antenna.

table 13. Maximized gain and F/B iteration for taper of 0.125 with a reflector length of 40.250 inches.

director 1 (inches)	gain (dBi)	F/B (dB)
36.500	15.156	12.144
36.625	15.241	12.282
36.750	15.324	12.524
36.875	15.404	12.909
37.000	15.481	13.494
37.125	15.548	14.377
37.250	15.598	15.734
37.375	15.613	17.948
37.500	15.562	22.124
37.625	15.386	36.110
37.750	14.982	23.407

table 14. Frequency response parameters for gain maximized antenna with a 0.125 taper.

frequency	gain (dBi)	F/B (dB)
142.5	15.409	12.006
143.0	15.481	12.794
143.5	15.546	13.879
144.0	15.597	15.457
144.5	15.613	17.948
145.0	15.554	22.586
145.5	15.353	36.116
146.0	14.902	22.227
146.5	14.014	14.837

table 15. Frequency response parameters for the "informational" antenna with a 0.125 taper.

frequency	gain (dBi)	F/B (dB)
142.5	15.473	12.544
143.0	15.540	13.618
143.5	15.593	15.171
144.0	15.613	17.609
144.5	15.562	22.124
145.0	15.370	36.809
145.5	14.925	22.548
146.0	14.035	14.849
146.5	12.286	9.831

table 16. Frequency response parameters for F/B maximized antenna with a 0.125 taper.

frequency	gain (dBi)	F/B (dB)
142.5	15.529	13.351
143.0	15.584	14.877
143.5	15.609	17.257
144.0	15.566	21.623
144.5	15.386	36.110
145.0	14.952	22.941
145.5	14.066	14.902
146.0	12.326	9.816
146.5	8.892	5.353

table 17. Maximized gain iteration for taper of 0.1875 with a reflector length of 40.375 inches.

director 1 (inches)	gain (dBi)	F/B (dB)
36.500	15.571	14.823
37.625	15.600	15.964
37.750	15.592	17.662
37.875	15.526	20.356
38.000	15.369	25.303
38.125	15.051	38.017
38.250	14.392	25.944
38.375	12.946	29.015
38.500	9.932	17.654

table 18. Maximized F/B iteration for taper of 0.1875 with a reflector length of 40.625 inches.

director 1 (inches)	gain (dBi)	F/B (dB)
37.500	15.563	15.153
37.625	15.596	16.300
37.750	15.590	17.990
37.875	15.524	20.638
38.000	15.367	25.413
38.125	15.051	39.162
38.250	14.397	27.566
38.375	12.935	20.931
38.500	9.828	17.129

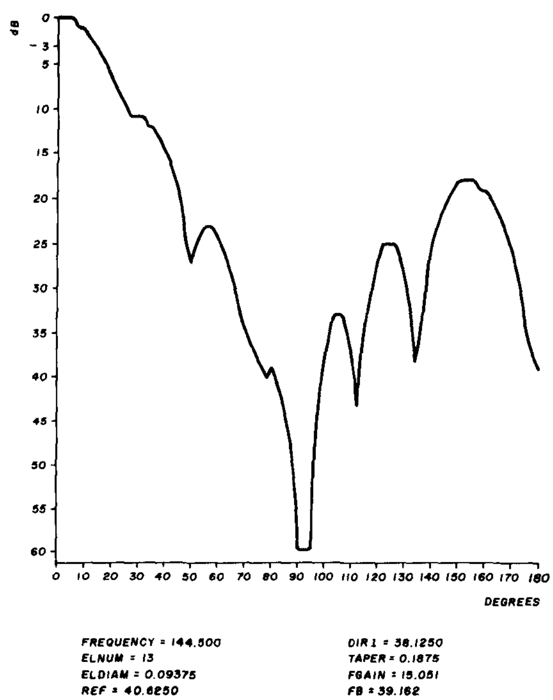

fig. 9. Plot of the 0.1875 taper, F/B optimized antenna.

tions of 15.613 dBi and 36.110 dB, respectively. These two antennas appear to be similiar because each optimized value lies directly next to an antenna with a director 1 length of 37.500 inches.

Frequency response calculations are provided for all three antennas. **Table 14** presents these calculations for the gain-optimized director length. This antenna's gain is virtually identical to its 0.0625 tapered counterpart, but its excellent F/B peak occurs far from the design frequency. **Table 15** presents the informational antenna's frequency response characteristics (**fig. 6**), which when compared to the gain optimized antenna, indicate a slightly lower gain and an excellent F/B ratio somewhat closer to the design frequency. **Table 16** presents the F/B optimized antenna's (**fig. 7**) frequency response calculations. They include a slightly lower gain than the gain optimized antenna and an excellent F/B at the design frequency and across the weak signal area. The designers' contentions for this antenna are largely supported by the computer iterations.

taper = 0.1875

This antenna, not presented by the designers, is a product of computer iteration. **Table 17** presents the gain optimized iteration for 15.600 dBi (**fig. 8**), and **table 18** presents the F/B optimized iteration for 39.162 dB (**fig. 9**). These two antennas are very dif-

table 19. Frequency response parameters for gain maximized antenna with a 0.1875 taper.

director 1 (inches)	gain (dBi)	F/B (dB)
142.5	15.389	12.283
143.0	15.458	12.861
143.5	15.520	13.608
144.0	15.571	14.602
144.5	15.600	15.964
145.0	15.590	17.904
145.5	15.517	20.841
146.0	15.342	25.865
146.5	14.994	39.126

table 20. Frequency response parameters for F/B maximized antenna with a 0.1875 taper.

frequency	gain (dBi)	F/B (dB)
142.5	15.593	15.293
143.0	15.588	16.761
143.5	15.536	20.048
144.0	15.379	25.109
144.5	15.051	39.162
145.0	14.356	28.320
'145.5	12.702	21.392
146.0	8.963	14.093
146.5	2.610	4.235

table 21. Maximized gain iteration for taper of 0.25 (linear) with a reflector length of 40.625 inches.

director 1 (inches)	gain (dBi)	F/B (dB)
37.625	15.479	14.317
37.750	15.535	14.871
37.875	15.570	15.619
38.000	15.572	16.625
38.125	15.529	17.973
38.250	15.424	19.764
38.375	15.224	22.055
38.500	14.848	24.520
38.625	14.060	25.524
38.750	12.332	22.659

table 22. Maximized F/B iteration for taper of 0.25 (linear) with a reflector length of 39.750 inches.

director 1 (inches)	gain (dBi)	F/B (dB)
37.500	15.420	12.601
37.625	15.471	13.028
37.750	15.506	13.627
37.875	15.520	14.466
38.000	15.506	15.652
38.125	15.454	17.378
38.250	15.348	20.059
38.375	15.154	24.938
38.500	14.784	43.202
38.625	14.006	25.211
38.750	12.344	17.292

table 23. Frequency response parameters for gain maximized antenna with a 0.25 (linear) taper.

frequency	gain (dBi)	F/B (dB)
142.5	15.421	13.148
143.0	15.525	13.779
143.5	15.536	14.474
144.0	15.570	15.420
144.5	15.572	16.625
145.0	15.525	18.120
145.5	15.407	19.816
146.0	15.175	21.245
146.5	14.713	21.394

table 24. Frequency response parameters for F/B maximized antenna with a 0.25 (linear) taper.

frequency	gain (dBi)	F/B (dB)
142.5	15.405	14.944
143.0	15.383	16.913
143.5	15.308	19.852
144.0	15,141	25.027
144.5	14.784	43.202
145.0	13.993	27.017
145.5	12.251	20.323
146.0	8.938	15.485
146.5	4.032	8.960

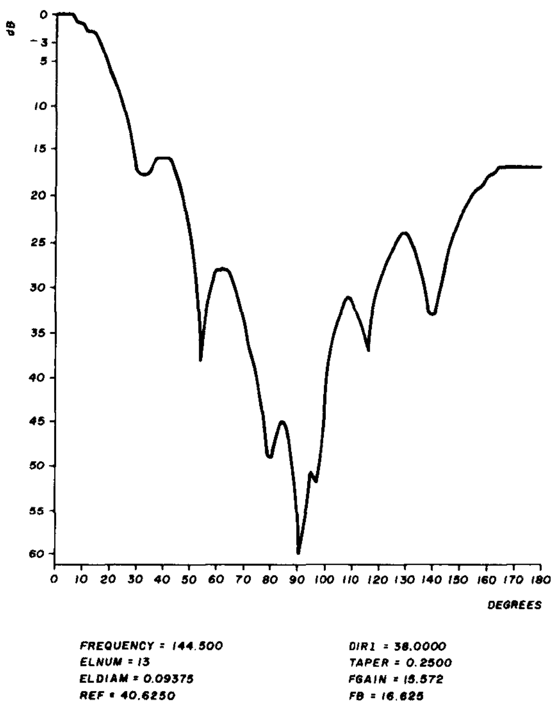

fig. 10. Plot of the 0.25 linear taper, gain optimized antenna.

ferent antennas. **Table 19** presents the gain optimized antenna's frequency response parameters, and **table 20** does likewise for the F/B optimized antenna. Both tables reflect gain figures that have fallen from comparable 0.125 taper antennas, but optimized F/B ratios have risen slightly.

taper = 0.25 (linear)

This antenna, also not presented by the designers, is another product of computer iteration. **Table 21** presents the gain optimized iteration for 15.572 dBi (**fig. 10**), and **table 22** does likewise for the F/B optimized value of 43.202 dB (**fig. 11**). These two antennas are significantly different. **Table 23** presents the gain optimized antenna's frequency response parameters, and **table 24** does likewise for the F/B optimized antenna. Within the 2-meter weak signal area, the gain optimized 0.25 (linear) antenna is virtually identical to its 0.1875 counterpart. This is also substantially true for the F/B optimized antennas for these two tapers.

taper = 0.25 (special)

The designers specified this antenna as having the best bandwidth characteristics of the three antennas they presented. No other rationale was specifically given for the somewhat unusual tapering. **Table 25**

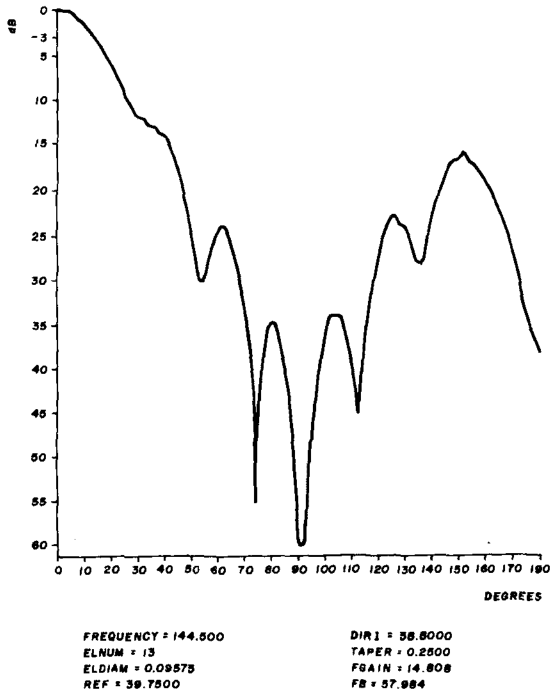

fig. 11. Plot of the 0.25 linear taper, F/B optimized antenna.

presents the gain optimized iteration for 15.573 dBi (**fig. 12**), and **table 26** does likewise for the F/B optimized iteration for 45.281 dB (**fig. 13**). **Table 27** presents the gain optimized antenna's frequency response parameters, and **table 28** does likewise for the F/B optimized antenna. When optimized for gain this antenna is quite similar to its 0.25 linear taper counterpart, but when optimized for F/B, this antenna has slightly more vectorial cancellation than its 0.25 linear taper counterpart.

computer iteration: summary

In terms of obtaining an optimized forward gain, any of the gain optimized antennas are satisfactory. The maximum calculated gain was realized with the 0.0625 taper antenna, one of the three antennas created solely by computer iteration. A very close second was the designers' 0.125 taper antenna. However, there is more to Yagi antenna selection than finding the absolute maximum forward gain among antennas that all have excellent forward gain.

A long Yagi should provide a sharp pattern for rejecting unwanted signals. This pattern should exist across the entire weak signal area. For the most part this is not the case with the F/B optimized antennas. With the possible exception of the zero taper Yagi, these Yagis are single frequency antennas, meaning

table 25. Maximized gain iteration for taper of 0.25 (special) with a reflector length of 40.5 inches.

director 1 (inches)	gain (dBi)	F/B (dB)
37.500	15.524	14.331
37.625	15.564	15.021
37.750	15.573	15.963
37.875	15.540	17.252
38.000	15.446	19.024
38.125	15.261	21.456
38.250	14.913	24.637
38.375	14.200	27.554
38.500	12.619	25.724
38.625	9.461	18.102
38.750	4.606	9.264

table 26. Maximized F/B iteration for taper of 0.25 (special) with a reflector length of 40.0 inches.

director 1 (inches)	gain (dBi)	F/B (dB)
37.500	15.527	13.709
37.625	15.555	14.421
37.750	15.556	15.420
37.875	15.518	16.836
38.000	15.423	18.905
38.125	15.239	22.129
38.250	14.894	28.033
38.375	14.194	45.281
38.500	12.676	26.475
38.625	9.693	18.000
38.750	5.115	9.767

table 27. Frequency response parameters for gain maximized antenna with a 0.25 (special) taper.

frequency	gain (dBi)	F/B (dB)
142.5	15.409	12.698
143.0	15.472	13.248
143.5	15.526	13.939
144.0	15.563	14.823
144.5	15.573	15.963
145.0	15.539	17.420
145.5	15.435	19.205
146.0	15.226	21.079
146.5	14.813	22.092

table 28. Frequency response parameters for F/B maximized antenna with a 0.25 (special) taper.

frequency	gain (dBi)	F/B (dB)
142.5	15.463	16.081
143.0	15.394	18.454
143.5	15.235	22.110
144.0	14.904	28.787
144.5	14.194	45.281
145.0	12.578	28.781
145.5	9.242	18.969
146.0	3.994	8.522
146.5	−2.248	−0.753

fig. 12. Plot of the 0.25 special taper, gain optimized antenna.

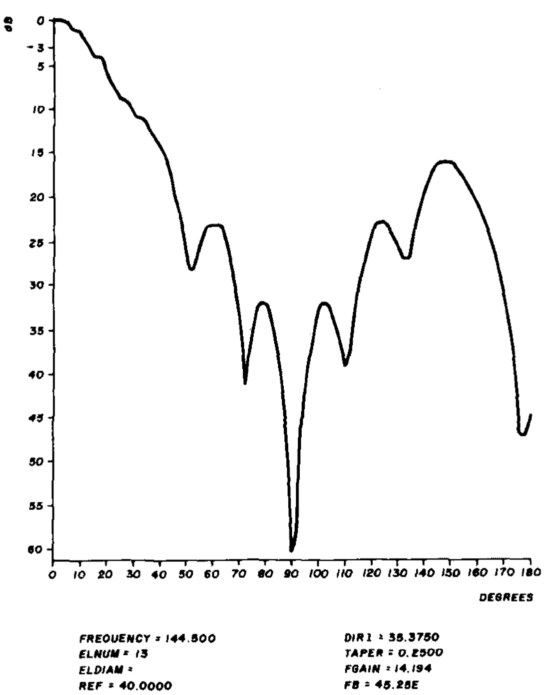

fig. 13. Plot of the 0.25 special taper, F/B optimized antenna.

their excellent F/B ratios are the result of single frequency vectorial cancellation. Their main lobes are broader and less clearly defined as compared to their gain optimized counterparts. While a high F/B is commendable, interfering or other unwanted signals do not always originate from the rear of a rotating antenna. A sharp main lobe peaks the desired signal and the Yagi's overall pattern reduces unwanted signals. For these reasons none of the F/B optimized Yagis would be the antenna of choice.

The zero, .0625, 0.125, and 0.1875 tapering procedures provided the highest optimized gains. In terms of the definition of the main lobe and the reduced amplitude of the first minor lobe, the 0.1875 gain optimized Yagi would be an antenna of choice. The calculated differences among these gain optimized Yagis are small and unlikely to be measurable in practice. While none of these Yagis has a notable F/B, their patterns are clean and have well defined nulls. The stacked Yagis at WB3BGU have a pattern similar to these single optimized Yagis. During the January 1984 VHF Sweepstakes, two important multipliers were worked in successive QSO's. One was from the front of the array (5-9), the other from the rear (5-1). A higher F/B would have resulted in the loss of the second multiplier. The array's clean pattern did noticeably reduce contest-level QRM.

Computer iteration has also helped underscore the use of even slight director tapering in long Yagis. Tapering up to approximately 0.55 degrees (0.125 inch at 144.5 MHz) appears to contribute to optimizing both gain and F/B ratios. Even when loops instead of rods are used for the reflector and driven element, a director taper of 0.125 inch (at 144.5 MHz) has been empirically found to provide maximum gain.[9]

It is also worth noting that in spite of the designers' claims to the contrary, tapered director Yagis require a longer first director than a comparable non-tapered Yagi. Arbitrarily tapering non-tapered director lengths that have been optimized for gain or F/B will result in measurably less gain and considerably less F/B. Tapering requires a distinct protocol for directors that when followed can result in a Yagi with superior performance levels.

Other problems with the designers' claims were found. Kmosko and Johnson made unusually high gain claims for their antennas. None of the iterations substantiated these claims. However, the designers do state that they never measured the actual gain of their antennas, but relied on a gain formula based on half power points. This inaccurate method always overstates forward gain. The measured gain for the NBS 3.2 wavelength antenna is 13.2 dBd. If based on the formula Kmosko and Johnson used, this same gain would be given as 15.53 dBd. It is also interesting to note that the designers stated they made careful pattern plots of what they found to be optimal designs.

It is hard to believe that a measurement as basic as actual forward gain would not have been made along with the rather tedious plots. Some of the problems attendant with reproducing the results the designers state for their antennas have also been mentioned by Reisert.[10]

An interesting comparison can be made between the NBS 3.2 wavelength Yagi and the above mentioned Kmosko-Johnson gain optimized Yagis. The NBS Yagi had a calculated forward gain of 15.20 dBi while the designers' Yagis averaged 15.60 dBi. The difference in gain is more than could be expected on the basis of the increased boom length (0.24 wavelengths). The variable parasitic element spacing used in the designers' antenna provided higher forward gain figures with a comparable minor lobe structure. The NBS Yagi provides the higher F/B, and measurably so. The 144 MHz operator concerned about F/B and desiring a shorter boom might opt for the NBS Yagi.

Computer iteration has been used to explore a large range of the larger total number of possibilities of the Kmosko-Johnson design. Selections made from among the twelve optimized antennas is a function of the user's needs and desires. There is no such thing as the single best antenna for everyone.

to be continued

The balance of this series will address individual VHF and UHF antenna designs. Next month I'll present new findings on the Greenblum antenna design approach, as illustrated by a long Yagi specifically designed for the 220 MHz band.

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ham radio

the high-performance, capacitively loaded dipole

Use distributed components for greater efficiency, higher gain

The capacitively loaded or "stretch" antenna is a wire dipole, one wavelength long, with each leg broken up periodically by equal value capacitors. The effect is to create a uniform current distribution over the entire length of the radiator without the use of phasing stubs or lossy inductors.

The stretch dipole, classified as a version of the Franklin antenna by E.A. Laport of RCA,[1] offers the following advantages:

- Low angle radiation as a vertical antenna.
- Higher efficiency (because its radiation resistance is at least twice as high as a quarter-wave monopole or dipole).
- Proximity to trees, towers, downspouts, and other antennas does not affect performance unless these structures are resonant at the operating frequency.
- Shape is not too important; the stretch dipole can be bent to fit into a confining area.
- The voltage and current are low compared to a quarter or half-wave antenna fed with the same power (input).
- The bidirectional pattern (figure-eight) is narrower, corresponding to higher gain.
- Sidelobes do not exist at the design frequency, because there is no phase reversal.
- It provides wideband performance.
- Deep nulls exist off the ends.

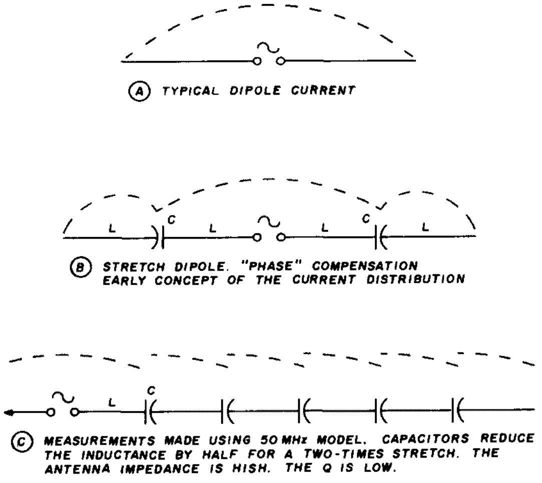

fig. 1. Examples of stretch dipoles and their respective current distributions.

antenna development

Forty years ago the technique of phase compensation in antennas was introduced.[1,2] Patents were applied for in both the United States and the United Kingdom. In 1960 F.J. "Dud" Charman, G6CJ, derived and tested antennas stretched to four times their normal length (**fig. 1**).[3] Recently W4FD and W4ATE produced the CCD design, which uses up to 48 sections and 46 capacitors.[4] VK5NN has built several of the twelve-section versions with a two- and three-times stretch.[5] The greater the stretch factor, the greater the gain because of the narrower figure-eight pattern.

The logic used in the design of this antenna stems from increasing the inductance of a radiator by a factor, (in my case, two), then reducing this inductance with capacitive reactance equal to half this amount. If twelve sections are used, each joined by a capacitor, each wire length is 360 degrees divided by 12, or 30 degrees long. Its inductance can be found by using published curves[2] or the formula:

$$L(\mu h) = 0.00508\,\ell\,[2.303 \log_{10}(4\,\ell/d - 0.75)]$$

where both ℓ and d are in inches.

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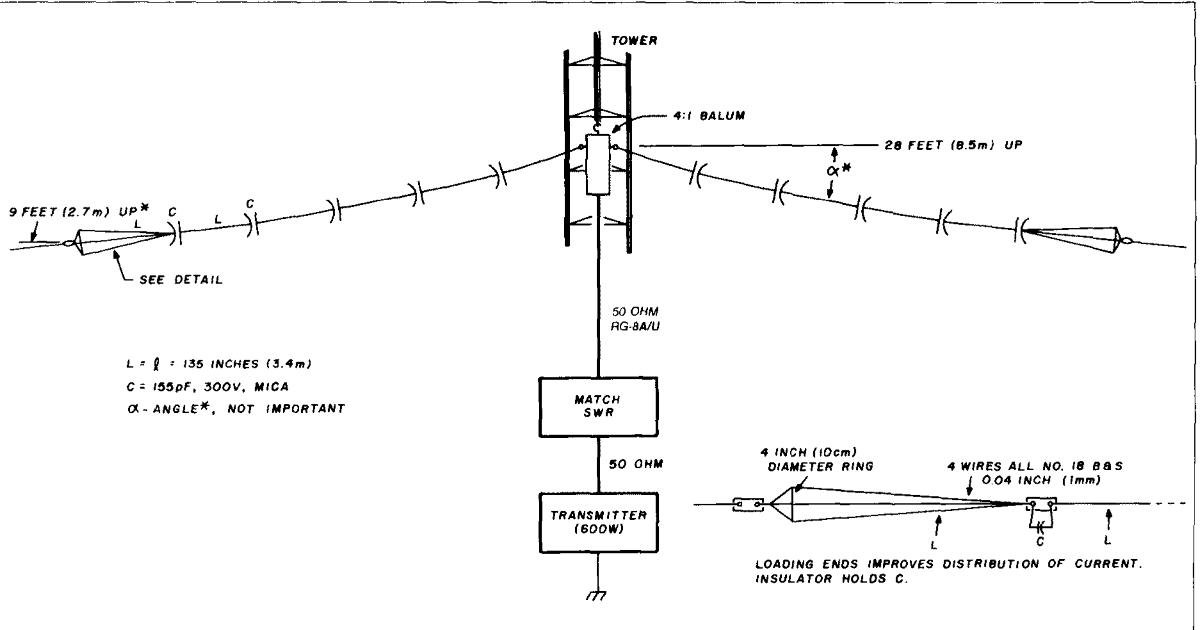

fig. 2. The 7 MHz stretch dipole measuring 132 feet in length can be supported from a tower in an inverted Vee configuration.

construction

Three different antennas were built. The first was designed for 40-meter operation at two-times stretch with an overall length of 132 feet (40.2 m) — see **fig. 2**. The second antenna, built to test the distribution of current and voltage, is also a two-times stretch dipole for operation on 6 meters. It was suspended at 6 feet (1.8 m) for convenience in measuring. (**Fig. 1C** shows the current distribution measured.) A GDO was coupled to a 300 ohm twin-lead, one wavelength feed via a 1:4 miniature balun. The current test meter used is shown in fig. 3. The third antenna was designed for 2-meter operation and features the same two-times stretch. Built using ceramic NPO 5 percent capacitors, this antenna is enclosed in 3/4 inch (2.5 cm) outer diameter thin-wall plastic conduit. End caps support the antenna within. A 3/4-inch "T" fitting at the center holds the third piece of conduit. An end-cap is drilled and terminals supplied for the 300 ohm twin-lead feeder, (see **fig. 4**).

For the lower frequency antennas, **fig. 5** gives values of section lengths in inches. Choose a frequency at the bottom of the graph. Where this value meets the "ℓ" line, read the wire length at the right edge of the chart. The wire size, plus or minus a size, is not critical, as the antenna is fairly wideband. Smaller diameters lower the chosen frequency slightly; No. 18 is a good choice. The capacitor values may be chosen for the same frequency by using the "C" line.

If the chosen frequency is off the graph, interpolation will give the desired values. For instance, for 144 MHz, use 14.4 MHz and divide the values by 10. For 7 MHz, use 70 and multiply by 10.

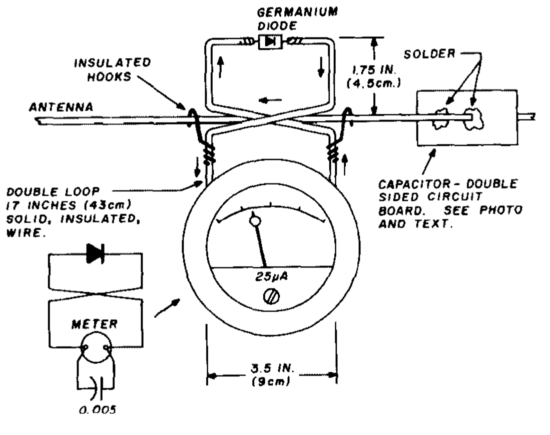

fig. 3. Current pickup used for 50 MHz measurements.

capacitors

The 40-meter full-wave dipole uses 300 volt mica capacitors. If micas are not available, double-sided circuit board makes a fine substitute. (So do the NPO ceramics.)

I measured a sheet, and it came to 18 pF per square inch. It is Polyclad-M, and 0.059-inch (1.5 mm) thick. This comes to 8.3 square inches, at 40 meters. Each piece would be, for example, 1.25 inches wide by 6.66 inches long (3.2 by 17 cm). The value should be approximately 150 pF. Measure any of this double-sided circuit board, because it can vary from type to type as much as 3 to 1. Solder the sections of wire to the

center of the boards, and again just in from the narrow dimension as shown in **fig. 3.**

I weatherproofed the capacitors using a mixture of one part resin (the kind used by violinists and sold in music stores) to two parts beeswax. This clings even in hot weather. It is best to heat the mixture (over low to moderate heat) over water — preferably in the top of a double boiler — and dip the assembled junctions into the solution when the temperature reaches approximately 150 degrees F (63 degrees C). Dip each piece for 5 seconds; remove; allow to drip and cool well away from the flame or burner, and then dip it

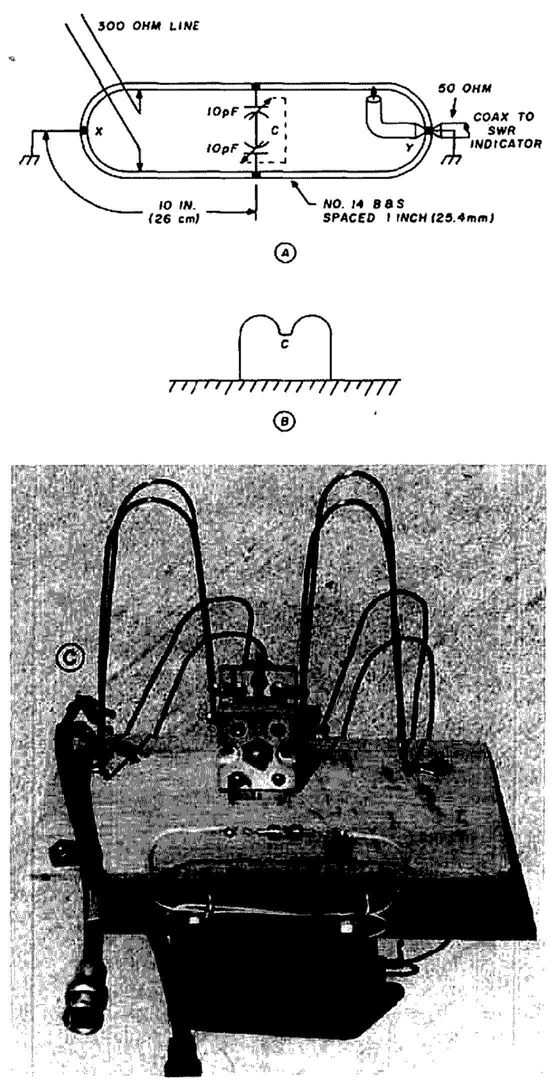

fig. 4. (A) Matching device transforms 50-ohm coaxial cable to 300-ohm open-wire feeders using a 40-inch (1.2 mm) conductor. Points X and Y are grounded and may be spaced 5 inches (13 cm) apart by bending the lines into inverted U's as illustrated in (B). The capacitor *C* is not grounded. (C) photo illustrates U-shaped matching device, capacitor, and current probe.

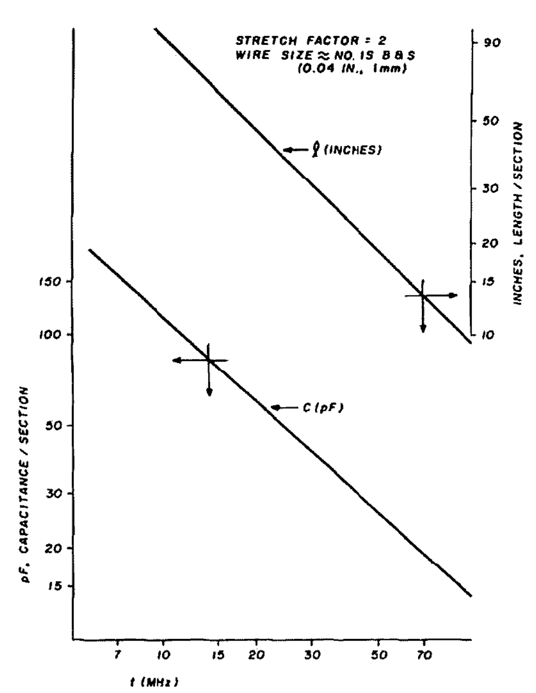

fig. 5. Chart can be used to determine section lengths and required capacitances.

again quickly, lowering it in and pulling it out almost immediately. This builds up the coating. This weatherproofing is used on the capacitors for all three antennas.

conclusion

I have experimented with a version of the phase-compensated wire antenna with a stretch factor of two. This two-times factor keeps the size practical from 7 MHz upward. Half of this antenna may be used as a vertical against ground-mounted radials or a counterpoise. The feed is higher impedance, resulting in lower heat losses in both the wire and ground. End loading helps develop uniform current at the top or ends in a full-wave horizontal or vertical dipole. These antennas may be used on all frequencies above the design band. The toroidal pattern, however, changes and minor lobes develop with increased frequencies due to changes in L and C reactances. A trans-match is a necessity.

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ham radio

remote-controlled 40, 80, and 160-meter vertical

70-foot irrigation pipe and switched L,C components provide 3-band coverage

A vertical antenna with an extensive ground system is effective on 40, 80, and 160 meters because it has a low angle of radiation; a horizontal antenna would have to be very high to achieve the same low radiation angle.

This article describes a vertical antenna system that I have designed, built, and tested. It operates over the entire 40, 80, and 160-meter bands and features remotely switched antenna base matching networks controlled from a simple box at the operating position.

The antenna is built from two 40-foot lengths of 4-inch diameter thin-wall irrigation pipe, available for approximately $35 per length. One length is left uncut; the other is cut into two pieces, one 30 feet long and one 10 feet long. The antenna is constructed from the 30 and 40 foot pieces joined together for a total height of 70 feet. The remaining 10-foot section is cut lengthwise, spread over the antenna sections, and clamped with four hose clamps to join the two main sections of the antenna.

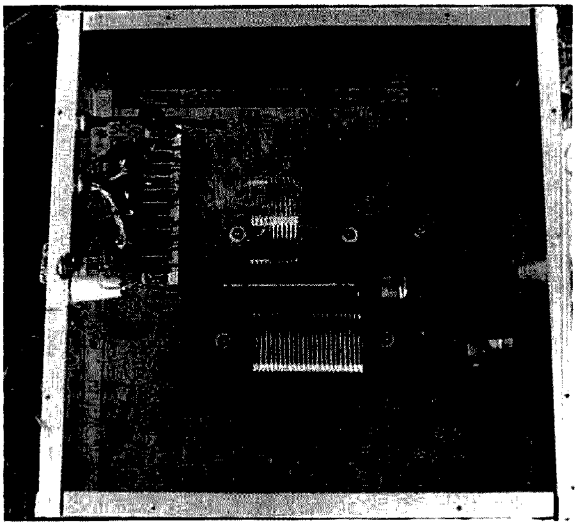

fig. 1. Large metal enclosure houses matching networks, Ledex solenoid and switch wafers while preserving high-Q of coils.

The bottom of the antenna is about a foot off the ground, bolted to a short piece of PVC pipe — positioned in the center of a 2-foot square of aluminum and sunk approximately one foot into the ground. The ends of the radial wires are bolted to the aluminum square. For efficient operation, about fifty 100-foot radial wires are used as a good ground mat. Guys are set at 10-foot intervals, with four nylon cords at each level. (S-shaped metal hooks can be used to fasten around the hose clamps and around guy thimbles, which in turn secure the nylon guys. The hook ends should be squeezed down so that they don't come loose from either the clamps or the thimbles. Black plastic tape wrapped around the nylon guy knots helps prevent the knots from working loose.)

installation

The antenna is light but quite flexible, so be sure to erect it on a calm day, with plenty of helpers. For safety's sake, check for nearby power lines that could be hit if the antenna gets away from you. I used a husky helper pulling a rope threaded through a pulley near the top of a temporary 25-foot mast to help raise the antenna. Another helper pushed from below (a temporary light A-frame might help here), and several helpers held nylon guys to prevent side sway. A bolt through the PVC pipe and near the bottom of the antenna provides a pivot point; I put a 2-foot piece of round wooden fence post inside the bottom of the antenna so that tightening the bolts there wouldn't squash the tubing.

matching networks

The 70-foot length of antenna is close to a quarter wavelength on 80 meters, about one-eighth wavelength on 160 meters, and about a half wavelength on 40 meters. It is therefore necessary to use matching networks at the base of the antenna on all three bands. The networks are remotely switched from the shack

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by means of a Ledex solenoid that turns a wafer switch, moving the shaft one position — a 45-degree rotation for each pulse. The switch wafers connect any one of the three networks and provide position information to illuminate appropriate indicator lamps in the shack. The five positions required by the three switch wafers coincide with the Ledex rotation positions.

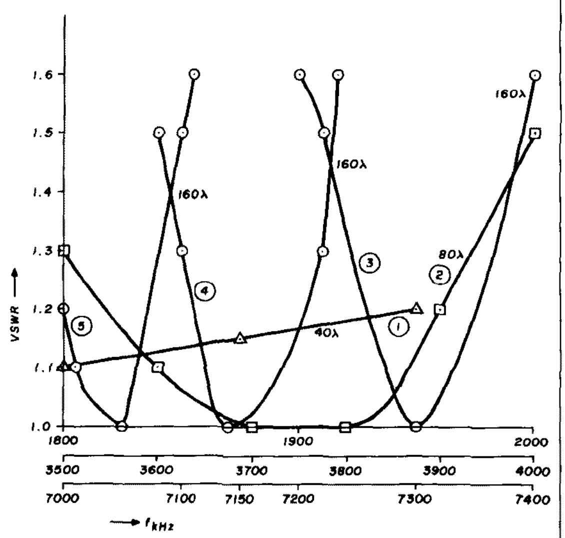


fig. 2. Five separate VSWR plots versus frequency of the 70-foot vertical and matching networks are illustrated with "1" for 40 meters, "2" for 80 meters, and "3" through "5" for the 160-meter band. Notice how flat both the 40 and 80-meter curves are.

The fixed component networks for 40 and 80 allow the antenna to be used over those entire bands without retuning and with a low VSWR. For 160 meters, the same kind of network is used over the entire band, but requires three coil taps that must be switched in to maintain low VSWRs for that band. The five switch positions are used as follows: one for 40, one for 80, and three for 160. In order to design the matching networks, it is necessary to obtain an estimate of the antenna base impedance for each band from handbook charts. The 70-foot antenna is inductive over the 80-meter band, with the resistive part of the impedance close to 50 ohms. It can be matched over the entire 80-meter band with nothing more than a series capacitor. For 40 and 160 meters, L networks are necessary.

I used a Smith chart to plot the antenna base impedance for the 80-meter band; that plot showed high VSWR at the high end of the band. Using a series capacitor changed the impedance plot so that the VSWR was always less than 2.0:1 over the entire band. Selection of the optimum size of series capacitor for these

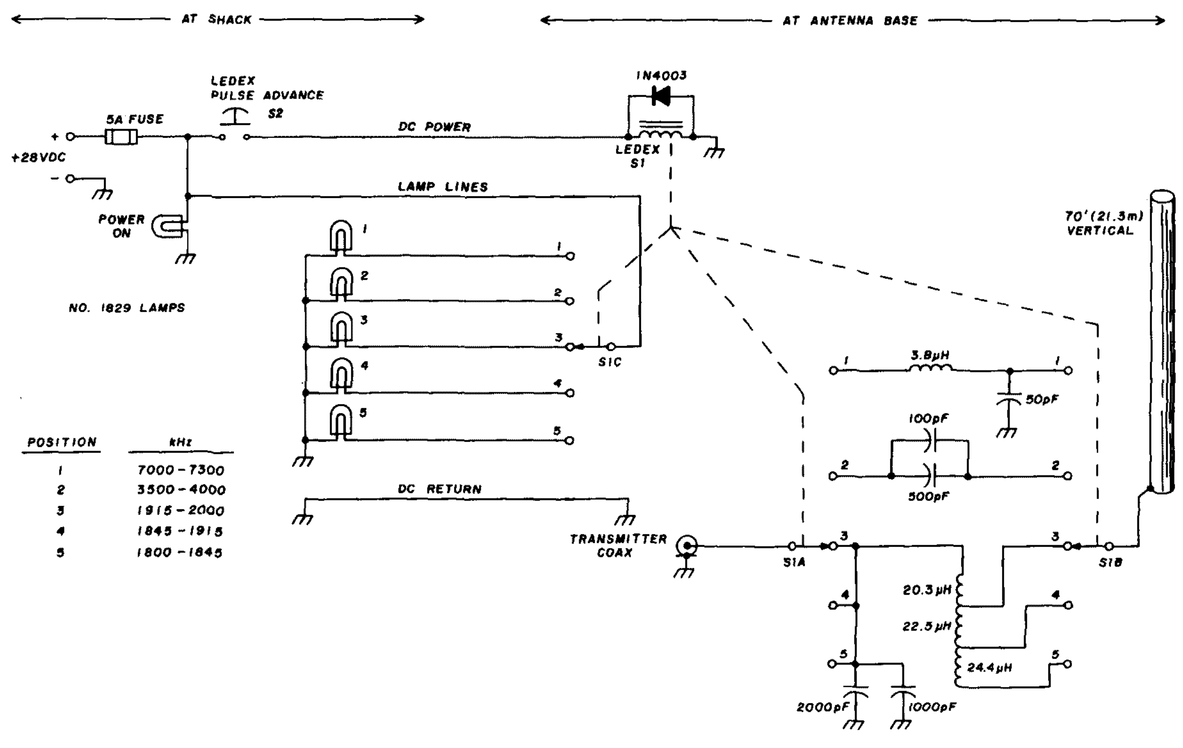


fig. 3. Schematic of control and RF circuit wiring.

Smith chart plots to give the least band edge VSWR provided an estimate of what size capacitor should be tried at the antenna base during field tests. Both the plots and the field tests showed that 600 pF was optimum. The 40 and 160-meter L networks were designed using standard handbook information.

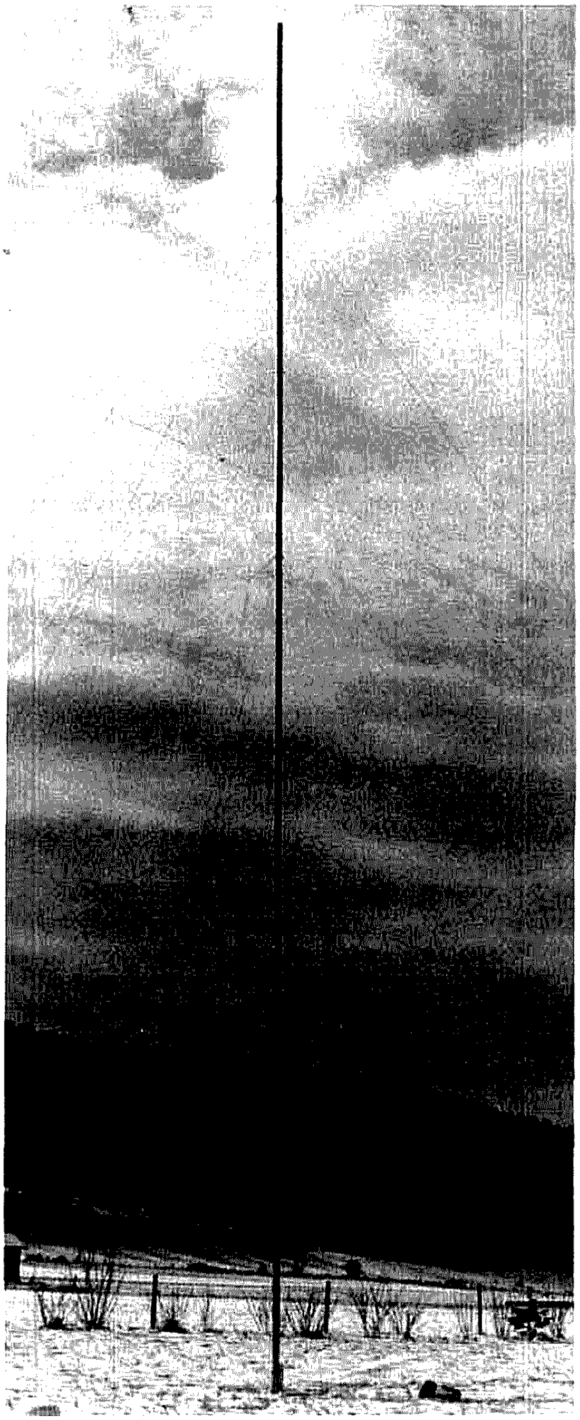


fig. 4. Four-inch irrigation pipe provides good mechanical and electrical performance for the 70-foot vertical.

network values

To refine the network values, I built a test network for each band and then tuned or changed each one for minimum VSWR over each band at the base of the antenna, using a small exciter and a VSWR meter. Next I measured the network component values and then built the final networks.

I made a final check of the VSWR at the base of the antenna using these final networks, and found that some pruning was still necessary. The tap locations on the 160-meter coil are quite critical. On my coil the tap for the 1800 to 1845 kHz segment was at the very end of the coil; the 1845 to 1915 kHz tap was about 1-1/2 turns from the end; and the 1915 to 2000 kHz tap was about 3 turns from the end. As **fig. 1** shows, a large metal enclosure houses the networks, Ledex, and switch. This box helps preserve the high Q of the coils by keeping metal surfaces some distance from the coils and also preserves essentially the same tuning for the L networks regardless of whether the box cover is on or off, so that you can tune with the cover off and still have the same tuning when the cover is replaced. As shown in the handbooks, the VSWR is lower back in the shack than out at the antenna, and in my case I find that for a 200-foot length of coax between antenna and shack, the maximum

fig. 5. Six separate sets of four guy wires are fastened to the vertical at 10 foot intervals using S-shaped metal hooks and hose clamps.

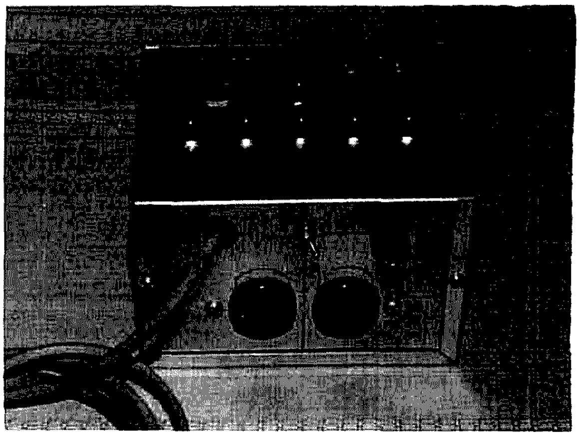


fig. 6. Simple control box allows for almost instantaneous no-tune band change.

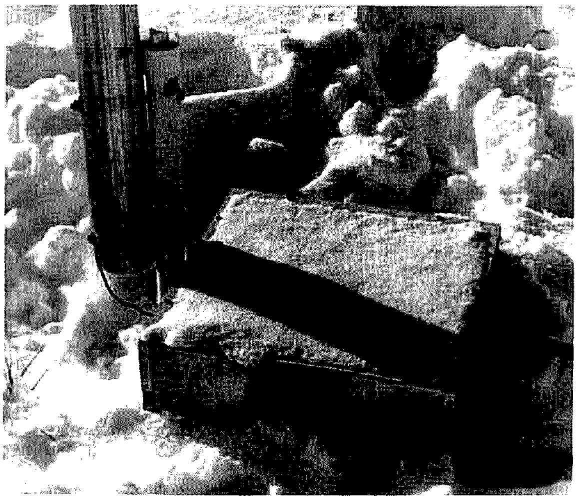


fig. 7. The remote control box should be located as close as possible to the base of the vertical.

VSWR at the shack is 1.6:1, as shown in **fig. 2.** I used surplus RG-17 coax cable to reduce losses over such a distance.

Fig. 3 provides the schematic of the networks and control system. **Fig. 4** shows the 70-foot vertical from a distance, while **fig. 5** is a view of the antenna from below. **Fig. 6** shows the control box, **fig. 7** the antenna base.

materials

L1 6 turns Barker and Williamson coil stock No. 3033
L2 23 turns Barker and Williamson coil stock No. 3033
Ledex S8210-025

For sources of Ledex, contact author. A 10-inch length of B-W coil stock No. 3033, sufficient for both L1 and L2 may be obtained from Amp Supply Co., P.O. Box 421, Twinsburg, Ohio 44087, for $12.50, plus $2.25 shipping and handling. The aluminum pipe may be obtained from any irrigation supply company.

ham radio

vertical phased arrays: part 6

Building the array and measuring performance

In this final article of my series on vertical phased arrays I will discuss some of the practical aspects of putting up an array — how to build it, how to construct networks, and how to take measurements. I will also address a few questions readers have raised about my previous articles.

siting elements: with respect to the world

Situating elements by eye can be deceiving. Having said this, I am absolutely certain that some will try it, nevertheless. Hopefully, you will discover any errors before a large radial ground system has been installed. Unlike adjusting the elements of a rotatable Yagi, adjusting the spacing of a ground-mounted vertical phased array is a major undertaking that may require *several weeks* of effort. If you know the variation from true north that your magnetic compass tells you is north, fine. Otherwise, the best way is to line up with the north star, Polaris. This star is easy to locate; the outer two stars outlining the dipper of the Big Dipper form a pointer to Polaris. I have a 4-square array whose major lobes are turned off of the desired directions because *I* failed to determine *the local* magnetic variation. Sources for this information include your local airport, any office of the FAA, or persons associated with private aviation. Determine whether the variation is east or west. Generally, this variation will be *west* for those located east of a line running through Chicago and Miami and *east* if located west of that line. For example, at New York City the variation is approximately 12 degrees west. This means that true north for a magnetic compass pointing at north is 12 degrees rotated clockwise toward the east. This variation from true north slowly changes with time; if your information is more than 10 years old, find a more recent source.

Since most of these arrays have half-power beam widths of 90 degrees or more, why be so concerned over a few degrees? For forward gain small errors in pointing do not matter much; we are more interested in the directions in which the beam should *not* be pointing. Just as with Yagis, it is far easier to determine the *direction of nulls* than maxima. This is important diagnostic information: to the extent that these are in the directions and reduced with respect to forward gain as predicted, we have a reliable validation of the design.

siting elements: within the array

Accurately locating the elements of an array, particularly if they are not to be in line, isn't as easy as it might appear. Getting the correct angles is the problem. Euclid had the right idea; three points not in a straight line uniquely define any triangle. Using wire with little or no stretch (steel or aluminum fence wire is excellent), carefully measure out three lengths, each equal to a side of any triangle that outlines all or part of your array. Join the ends, and with two helpers, pull the wires taut, you'll have three points accurately located with respect to each other. If your array is triangular, you're all set. If it's a 4-square, you have only to locate the fourth element with the same wire triangle by turning it over on its diagonal. Triple-check

By Forrest Gehrke, K2BT, 75 Crestview Road, Mountain Lakes, New Jersey 07046

table 1. Single 80-meter element tubing requirements.

quantity	length	diameter	wall	cumulative height
3	10′ (3.05 m)	1-1/2″ (3.81 cm)	0.125″ (3.18 mm)	30′ (9.14 m)
1	10′ (3.05 m)	1-1/4″ (3.18 cm)	0.125″ (3.18 mm)	39′4-1/2″ (12.00 m)
1	8′ (2.44 m)	1″ (2.54 cm)	0.057″ (1.45 mm)	46′8″ (14.22 m)
1	8′ (2.44 m)	7/8″ (2.22 cm)	0.057″ (1.45 mm)	54′2″ (16.51 m)
1	8′ (2.44 m)	3/4″ (1.91 cm)	0.049″ (1.24 mm)	61′8″ (18.80 m)
1	4′ (1.22 m)	1/2″ (1.27 cm)	0.049″ (1.24 mm)	63′6″ (19.35 m)

Additional material requirements for a single element.

2	15″ (0.38 m)	1-1/4″ (3.18 cm)	0.125″ (3.18 mm)	mating inserts
1	24″ (0.61 m)	1-1/4″ (3.18 cm)	0.125″ (3.18 mm)	extender
1	18″ (0.46 m)	7/8″ (2.22 cm)	0.049″ (1.24 mm)	reinforcement
7	S.S. helical hose clamps approximately 2 inches (5.08 cm) OD			
9	S.S. 1/4″ - 20 1/2″ screws			
8	S.S. 8-32 1/2″ screws			
1	0.250″ (6.36 mm) female quick disconnect terminal			
1	0.250″ (6.36 mm) male quick disconnect terminal			
1	SO-238 UHF female terminal			
12″ (30.5 cm) flat tinned copper braid				
500′ (152.4 m) 1/8″ (3.18 mm) nylon woven cord				
7800′ (2377 m) PVC insulated No. 24 solid copper wire (100 0.3 wavelength radials)				

everything to be sure, because array element layout is one of the few *physical* items under your complete control among the factors determining array symmetry. In prior articles I showed that electronic beam switching requires every element to operate *identically* in each of the *different* electrical positions of the array. This is a severe requirement; the best we can hope for is to get within 5 percent of meeting it, realizing that reaching within 10 percent results in a significant loss in F/B performance.

For those who may want to check array patterns, I have observed that reception of a 1-watt signal source located between a 1/4 to 1-mile distance is consistent with the pattern that is seen at the vertical angle of maximum radiation (but without QSB). However, at 20 miles this is no longer true because high vertical angle reflections predominate, sometimes so strongly that a positive F/B is seen.

monopole construction

After much experimentation with a variety of ways to put together tubular quarter-wave length ground-mounted 80-meter vertical elements, I hit upon a method of construction which has held up for over six years. It's relatively inexpensive, but has withstood the rigors of northeastern winters, including icing followed by 80 MPH winds. After failures with lighter designs I decided that, at least for 80 meters, any tubular construction must be able to withstand being raised in one piece. If a vertical can withstand such stress, then it should also be able to survive high winds, icing, and even the temporary loss of one or two of its nine guys.

Table 1 lists dimensions of aluminum tubing that, when assembled into a quarter-wave element, will meet this criterion. Included with the table is a complete list of materials for a single element. If care is taken not to raise the antenna abruptly, it will stand tall and straight — despite all appearances to the contrary — as it is brought upright.

All tubing will telescope into its next larger diameter mating member except the lower 1 1/2 inch (3.81 cm) diameter lengths. For two of these lengths, a 15-inch (38.1 cm) section of 1 1/4 inch (3.18 cm) diameter 0.125 inch (3.18 mm) wall tubing is bolted (using three 1/4 - 20 screws) at one end with 7 1/2 inches (19.05 cm) protruding, forming a mating junction with the next lower identical diameter tubing. The 1-inch (2.54 cm) diameter length of tubing requires a 15-inch (38.1 cm) length of 7/8 inch (2.22 cm) diameter tubing to be inserted for its entire length at the lower end to act as reinforcement because of the abrupt change in wall thickness at this junction. All lighter tubing is drilled and tapped for stainless steel 8-32 screws at two places spaced about 5 inches (12.7 cm) apart, at junctions. This is necessary to prevent the development of intermittent continuity after a few months due to wind vibration. The tubing, having little weight in this part of the vertical, cannot be depended upon to maintain good contact by gravity.

This element will resonate at approximately 3800 kHz. Inevitably, multiple elements will not resonate at precisely the same frequency even though they are identical in physical length. For exact matching of resonant frequencies, a 2 foot (61 cm) length of the 1 1/4 inch (3.18 cm) diameter 0.125 inch (3.18 mm) wall

tubing is used at the bottom of the vertical. This piece has tapped holes every 2 inches (5 cm) for a stainless steel 1/4 inch-20 screw, which determines the amount of its length that can be inserted into the bottom of the vertical. This may be adjusted as measurements dictate.

Flat braid [approximately 12 inches (30.5 cm)] is doubled, a 0.250 inch (6.36 mm) female quick disconnect terminal soldered at one end, and clamped to the bottom of the vertical with a helical hose clamp. I wrap PVC electrical tape around this to keep the doubled braid together. This makes a flexible, low inductance connection to the feeder. The coax termination is an SO-238 UHF female connector to which is soldered a male 0.250 inch (6.36 mm) quick disconnect terminal. The reason for these terminals will become quite obvious as measurements begin.

Glass bottles, corked to prevent accumulation of rain, may be used as standoff insulators for the verticals, since the necks happen to fit within the element base.

guy wires

Three sets of three guys, one set every 16 feet (4.88 m) from the base, are connected by two hose clamps at each attachment point. One clamp acts as a backstop for the clamp immediately above it, which clamps around the nylon guys. The nylon guy ends are tied with their own guy and also with one of the adjacent guys as additional insurance (falling tree branches can tear away the first tie but the fall, once arrested, seldom takes out the second tie). The attachment areas are waterproofed with PVC tape.

An element is raised by threading one of the three middle guys (usually made longer than those adjacent, specifically for this purpose) through a pulley which may be as low as 35 feet (10.7 m) from the ground. Since my array is among trees, I chose one to serve as a ginpole — which, of course, requires a real ginpole if you have no trees. Identify all guys with their ground anchor location, and lay them out so that no crossovers will be necessary later. During raising, the two remaining middle guys should be controlled by helpers to restrain the element from moving to the right or left, and as it arrives near the vertical position, to restrain it from continuing in the direction of the raising pulley. Don't forget to instruct your helpers in this latter point; more than one vertical has been successfully raised, only to continue unrestrained on its path to an inglorious end as it passes the upright position!

I've found that 1/8 inch (3 mm) diameter white woven nylon cord (sometimes called parachute shroud) is an economical, strong, long life material for guy stays. This is available at K-Mart stores in 50- and 100-foot (15 and 30 meter) lengths. I have some still in use after six years. The same cannot be said for polypropylene rope. Even 1/4 inch (6 mm) UV resistant material will fail in just two years.

radial systems

Installing radial systems is the dog work of building a low band array. It is also where the payoff — which too few Amateurs collect — is. There are two benefits to be gained with an extensive radial system: low losses and, more importantly, a low vertical radiation angle. But there's no free lunch: forget the loose talk you've heard on the bands about the benefit of water tables a foot below the surface, or being located over high conductivity earth. Pure water is a very good insulator, and most fresh water is too. And "high earth conductivity" is relative; it is *very* poor compared to the conductivity of copper. (For a perspective, see **table 2.**) Metal stakes in the ground at the base of your vertical give you good lightning protection, but *not* a good ground plane. Nor are there any rediscovered long-lost shortcuts; forty-eight radials raised a few feet above the ground won't provide any more efficiency than the same number on the ground. Undoubtedly, the best of all worlds would be an island surrounded by seawater, but for the near-field we'd *still* want an extensive copper radial system.

table 2. Conductivity comparisons.[1]

low conductivity earth	0.0005	mho/meter
high conductivity earth	0.03	mho/meter
sea water	5.00	mho/meter
copper	58,000,000.00	mho/meter

The best indicator of a good ground plane is how close the resistive component of the radiator's apparent self-impedance is to its theoretical resistance. The factors affecting theoretical radiation resistance are the electrical length and effective radius of the element, assuming a uniform cross-section monopole. Top hats, loading coils, and other means of shortening are also amenable to calculation, though the mathematics in some cases is more complicated.[2] For quarter-wave radiators this value is approximately 36 ohms.

In practice, there's another way to make this determination for any radiator without knowing the theoretical radiation resistance. It is the kind of analysis we usually wind up doing anyhow. Lay out radials, say ten at a time, distributed equally in all directions, taking measurements of the radiator's apparent self-impedance for each group of added radials. Plot these points on a graph as in **fig. 1** (open circuit any other elements of the array to avoid coupling). You will find that each lot of radials has less effect upon radiation

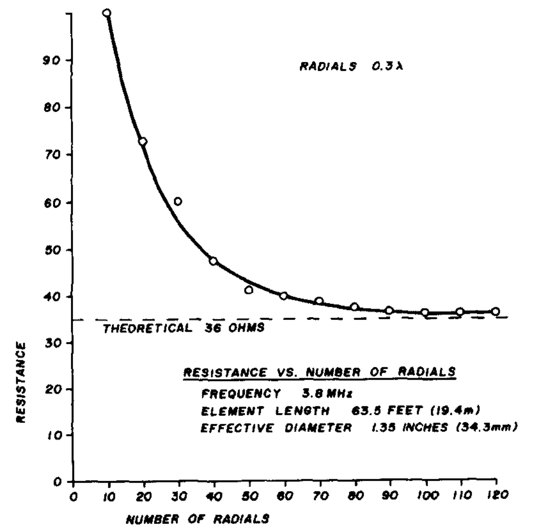


fig. 1. Input resistance of vertical antenna (includes all losses) versus number of radials.

resistance than the previous lot. Assuming radial lengths of a quarter-wave or more, after about 100 radials, the reduction in resistance with each added lot of radials is almost constant, but becomes vanishingly small (approximately 0.1 ohm). You'll notice that the curve on your graph has begun to flatten out; if you fit a French curve to this plot, you'll see that the plotted curve will nearly meet some horizontal line. In scientific terms, this is described as having become *asymptotic* with the theoretical resistance of the radiator; the horizontal line is a prediction of that theoretical value. (I am assuming negligible radiator element circuit losses. This is a fairly safe assumption for aluminum monopoles, but less safe if loading coils are present. More rigorously, the plot is becoming asymptotic with theoretical *radiation* resistance *plus* radiator circuit resistance.) Put another way, you've reached the point of diminishing returns. Although that point is self-definable, most experimenters would agree that it begins at the knee of that curve — i.e., at about *50* quarter-wave radials.

An aside to single vertical users accustomed to rating an antenna's merits according to VSWR readings: don't misinterpret an increase in VSWR as a negative indication when adding radials. Assuming VSWR is 1:1 with 50-ohm coax to a quarter-wave vertical, an appreciable ratio of output power (approximately 28 percent) is being used to heat the ground around the radiator. A higher VSWR after adding radials is desirable.

How should the radials be laid? This depends upon your personal aesthetics and also upon how the area occupied by the array is used. If you need to bury the radials (see "Build a Simple Wire Plow," page 107 — **Editor**), some form of protection against corrosion, such as PVC insulation or enamel coating, is necessary. Don't bury them too deeply; the closer radials are to the surface, the more effective they are. Some Amateurs have laid them flat on the surface and let grass cover them so well that limited traffic and even lawnmowers can be allowed.

What about wire size? I often hear people talk about laying No. 6 or No. 8 BS gauge radials. Unless you have to protect your system from farm animal traffic, this is calling for a hawser when thread will do. Consider: 1000 watts output to a single vertical with 25 radials (and assuming an apparent radiation resistance of 50 ohms). Each radial will be carrying all of 179 milliamperes, and *that* only near the base of the vertical.[3] Given a reasonable number of radials, a base current measured in amperes is divided into individual radial currents of milliamperes, and even this small current rapidly *decreases as we move* away from the base of the radiator. In the absence of concern about possible fragility, the wire size may be quite small.

Many articles in Amateur publications have suggested using steel fence wire or steel mesh as an economical substitute for copper. *Don't.* Unless the material has cost you nothing, and your labor is worth nothing, and you plan to abandon the antenna in less than a year, forget it. In just a matter of months, galvanizing — if present at all — is penetrated and corrosion proceeds. Steel, being magnetic, has a high permeability, making for a skin effect much thinner than copper when carrying RF current. Iron oxides are lossy semi-conductors. The thin skin effect, combined with a lossy surface, results in a wire which conducts *nearly zero* RF current long before it fails to conduct DC or has lost physical integrity — which takes only about three years after installation, and much less time if the system is buried. My first radial system consisted of a combination of *aluminum*, steel, and copper radials. It took a couple of years to solve the mystery of a slow but continually rising self-impedance of some of the elements in the array. In my efforts to overcome this rise, I compounded the mystery by adding more radials, which at first were — *more steel radials!*

array operation measurements

So you've constructed an array. Now you'd like to see how well it works. On-the-air tests are understandably at the top of your list. Unfortunately, this is not likely to be a good proof test of proper drive conditions, primarily because these arrays *want* to work and may show fair performance despite being poorly driven. This is almost always true for gain

characteristics, and during some propagation conditions may even apply to F/B. So continue these tests, but give some thought to an old antenna man's advice, said to have been first enunciated during the period of Maunder's Minimum: "One swallowe prouveth not that summer is neare."*

A much more definitive test is a measurement of element currents in each of the array directions. Assuming you have designed the feed network for a 1:1 VSWR, then element base current amplitudes measured within ±5 percent of design values and an array VSWR no greater than 1.15:1 in any direction is almost complete proof that drive is in the proper range, including current phase angles. Measurement of element current amplitude *and* phase is, of course, the ultimate test for drive conditons. A wideband dual-trace CRT is needed for this test, and since this equipment is quite expensive, may be beyond the reach of most Amateurs unless it can be borrowed. On the chance you have access to this equipment, a method is described in the next section of this article.

Measurement of current amplitude, a must for any serious array builder, is quite easy to do. At the lower frequencies a high degree of absolute accuracy is not difficult to achieve, but good linearity is really all that is necessary. For example, if actual current is doubled (power multiplied by 4), does the reading double? For this purpose the meter readings might just as well go from 1 ampere to 2 amperes, or 400 mA to 800 mA; we are more interested in good linearity of readings than in absolute value because phased array design considerations are concerned only with element current amplitude ratios.

Fig. 2 shows a schematic of an RF ammeter (**photo 1**). The basic meter movement may be anything up to 1 milliampere. I use germanium diodes for the rectifier because of their low turn-on voltage. This simple design works well for absolute accuracy and linearity up to 14 MHz. Low capacitance of the pickup coil to RF line is important and is increasingly so as frequency is raised. For this reason we want a high permeability factor for the toroidal core (for least number of coil turns). A Faraday shield will provide even more isolation, but this additional protection is not necessary at lower frequencies.

Since this ammeter is easy to duplicate, you may find it useful to have one for each element of the array because the efficiency of data collection is considerably improved. Note the use of quick disconnects for element terminals and associated measurement devices.

*From 1645 to 1715 there were no observable sunspots, and no Northern Lights. (Imagine a 70-year period in which the 10-meter band never opens and the 20-meter band is only so-so during the day, and dead at night!) A British astronomer, E.W. Maunder, in 1895, was the first to call attention to this strange behavior of the sun.

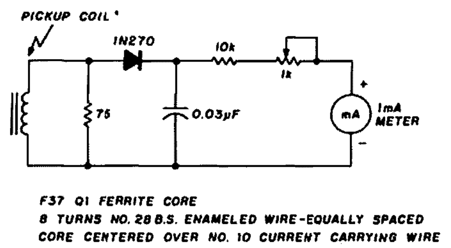
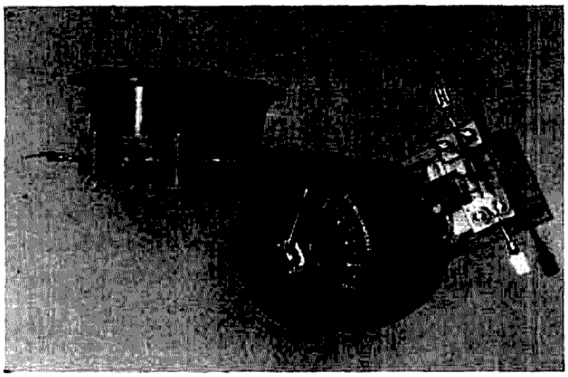


fig. 2. RF current probe circuit midscale reading corresponds to 1 ampere.



Solidly constructed RF ammeter uses quick connect/disconnect terminals.

dual-trace CRT measurements

Measurement of RF current phase angles involves an instantaneous comparison of sinusoidal currents at the bases of the elements. Since the elements are widely separated physically (and distance is proportional to phase), we must take special precautions to be sure we are really observing events in time coincidence. One way is to make these observations at another location of our own choosing in such a manner that all events have been equally delayed. Though we will see events at some time later than they occurred, they will be in time coincidence. Identical coax lines will meet this requirement nicely. Obviously the pickup coils for sampling base currents also must be alike, with the further proviso that the terminations of these lines must be alike, resistive, and match the characteristic impedance of the lines. (Most dual-trace CRTs provide 50-ohm inputs.) The line length chosen should be long enough to allow measurements of the most widely separated elements. I am sure you've anticipated my comment that the assurance of identical electrical length is not provided by a tape measure. Furthermore, to help ensure line identity for other characteristics, it would be a good idea to cut these

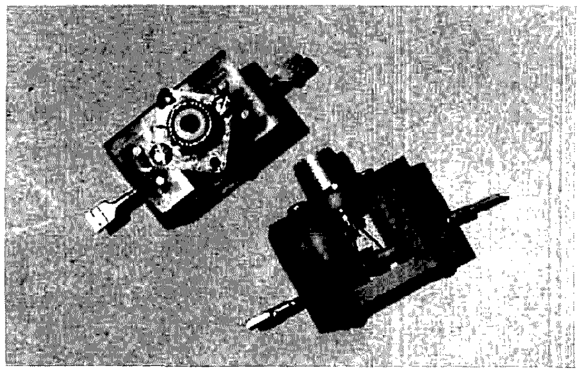
lines from a single piece of cable. (For an enlightening discussion on the variations that may be found in coaxial cable, see Bill Orr's "Ham Radio Techniques" in the January, 1984, issue.)

Photo 2 shows the construction of a pickup coil fixture. This is essentially the same as that used for an ammeter, but without the rectifier and filtering circuit. More attention must be paid to ensure that the pickup circuits have identical RF characteristics, however. We can test this by connecting both pickups in series as sensors in the same RF current circuit, preferably a resistive load. The amplitude and phase of the displayed current waveforms should be identical and should remain so when positions of pickups are interchanged. If reasonable care has been taken, the phase difference due to pickup coils should not exceed 2 or 3 degrees. Any small difference may be corrected for *a particular band* by connecting a mica capacitor (5-25 pF) across that pickup coil which *lags* in phase. My guess is that toroid core material variations are the probable cause of any slight differences. It is possible that substitution of another sample of the same core might also work; I did not investigate this.

In my experience with this measurement technique, I found it to be the most productive method for fine tuning a feed network to get the last bit of F/B performance improvement. Because both phase and amplitude changes are displayed, a much more rapid and intelligent analysis of cause and effect is possible. This comment applies with ever-increasing emphasis with the number of elements in the array. Lacking this capability, my advice is, "If it ain't broke, don't fix it!"; this can be as frustrating as attempting to adjust the color matrix board of a TV receiver without a crosshatch generator. The fact is that if self and mutual impedances are accurately read and used for the design and careful construction of the feed network, the array will be operating very close to, if not exactly at, optimum. The few adjustments determined with the dual-trace CRT are surprisingly miniscule "tweaks." Although the effect on F/B can be quite marked, for example, improving F/B from −20 dB to −30 dB, don't be carried away by these numbers: the effective frequency range over which this occurs is extremely narrow.

alternate methods of phase measurement

One would expect that considering its importance in antenna applications, measurement of phase angles at lower RF frequencies would have received more attention in the Amateur press than it has. A survey of Amateur publications did not yield much except one very interesting article directly applicable to phased array applications.[4] While the author's concept is ingenious and well chosen, it used a differential phase



Pickup coils provide signals for dual trace scope.

angle readout that was analog rather than digital. Considering the tremendous advances in semiconductor technology during the ten years since the article appeared, a digital readout should be possible. I hope that I may interest some enterprising experimenter to take up this challenge.

network construction

Several readers have commented that although the no-compromise advantage of 4-terminal feed networks is obvious, and that matrix algebra for the modular design of the networks is a powerful tool, they had apprehensions about how to turn the mathematics into working hardware. Some, familiar with the Pi networks seen in linear amplifiers, were discouraged by visions of the need for the same size of components and the cost of construction. Still others thought the networks might be difficult to tune.

None of these concerns are justified. Construction is actually quite simple, and with an accurate noise bridge,[5] tuning is easy. Tuning with an impedance bridge is done in the same step-by-step manner as is the design, and one of the prime advantages of this method is that it allows tuning to be done at the design frequency. The impedance levels of these networks are low, being generally in the 35- to 125-ohm range. Since each network chain is designed to appear *resistive* at its input and deals with only a portion of the transmitter power, voltages are seldom above 300 volts, even when driven by kilowatt linears. For example, I use postage stamp size mica capacitors extensively (500 volt rating) and I've yet to have one fail. Where high impedances are encountered, for instance, with elements requiring very little direct drive because of drive coupled from other elements, the current is so low that high voltage is not developed.

Photo 3 illustrates the simplicity and small component sizes. This is the feed network for my 80-meter 4-square array.[6] It is built into a 3 × 6 × 8 inch (7.6 × 15.2 × 20.32 cm) box on PC board, with each network chain individually removable. This takes the

place of other feed methods which require 130 feet (39 meters) or more of coax for this band. At today's prices for good quality 50-ohm coax, this alone should be the deciding factor, without even considering the superior technical merits of 4-terminal feed networks.

As can be seen from the photograph, small 100 pF air variable capacitors are used as trimmers. Mica capacitors, singly or paralleled, using their color coded values, are chosen to make the required network capacitance fall in the middle of the trimmer range. All inductors carrying significant current (and these tend to fall between 0.5 μH to 5 μH) are air core using No. 10 or No. 12 B.S. enamelled copper. I wind these on 1-inch (2.54 cm) diameter wooden dowels, letting them spring up to a slightly increased diameter. Inductances significantly above 5 μH are wound on powdered iron toroids using No. 18 B.S. wire. Using single layer charts for the air wound coils or toroid core manufacturer charts, all inductors are wound to be well above the inductance required. A grid dip meter, together with a known capacitance, is used to trim the inductors to slightly above the required values (5 to 10 percent). The network is constructed with these components and is completed with the exception that no network interconnections are made nor are any connections made to the *shack line* coax terminal.

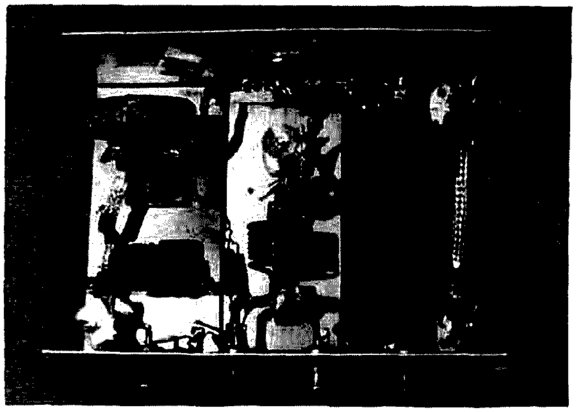
tune-up procedure

We have to choose which direction the impedance bridge will look into the network for tuning it. Each network chain was designed to transform a complex impedance to a pure resistance. Since it is much easier to duplicate resistances than complex impedances, this usually determines the choice. Assuming this case, consider a cascaded network consisting of a shunt L-match followed by a Pi circuit, which is the typical chain from the feeder of the −180 degree phased element of a 4-square array. For example, from Part 4[6] of this series, the input impedances as seen at the various points of a network chain are reproduced for element No. 4 of a 4-square array:

element No. 4 driving-point impedance	63.4 + j47.5
input impedance to 100 degree length feeder	22.73 − j11.37
input impedance to L-match	114.83 + j0
input impedance to Pi circuit	114.83 + j0

You will recall the reason for the Pi circuit was for phase matching, not impedance matching; therefore, its input impedance is the same as the impedance to the shunt L-match.

The impedance bridge unknown terminal is connected to the coax terminal, to which the element feeder would normally be connected. The resistor simulating the input resistance of the L-match (115 ohms is ok) is temporarily soldered to the input of the L-match at the shunt arm and ground (i.e., at the connection point normally going to the Pi circuit). Since we have chosen to measure conditions in the *reverse* direction through this network, we must consider what we want the bridge to "see" and set it accordingly. At first glance, it might seem reasonable to expect this to be *equivalent* to the impedance as seen at the input to the element feeder. Not so; this is not a bilateral case. Instead we should expect to see the *conjugate* of this impedance: i.e., 22.73 + j11.37. If the impedance bridge reads parallel circuit equivalents, these must be calculated and the bridge set accordingly. If the reactance is beyond the basic bridge range, an appropriate extender must be used. The tuning procedure is simplicity itself; without touching the bridge settings, adjust the L-match shunt trimmer for minimum detector output (being sure this minimum is within the trimmer range). If this is the normal impedance bridge null, the adjustment is complete. More likely, it is merely a minimum. Begin spreading out one of the outer turns of the inductor and readjust the trimmer. Since the null is sharp and deep, use care in spreading coil turns to be sure you have not passed through the null. (I use the tapered end of a pencil for this.) When the L-match is tuned, move the simulating resistor to the input of the Pi circuit (the point normally connected to the shack coax line terminal). Install an interconnection between the L-match and Pi circuit. With the bridge remaining connected and set as before, adjust the Pi circuit trimmers for minimum detector output and then reduce inductance by turn spreading. Since the Pi circuit has three interdependent adjustments, be sure to recheck the other two with each tuning change. The two trimmers should end up in approximately the same part of their range, assuming the fixed padders are similar. Since the tune-up of the Pi circuit is done with the bridge looking into the L-match, a separate procedure for integrating the two networks is unnecessary. Remove the resistor, but



Matching network for four square vertical array measures 3 × 6 × 8 inches (7.6 × 15.2 × 20.32 cm).

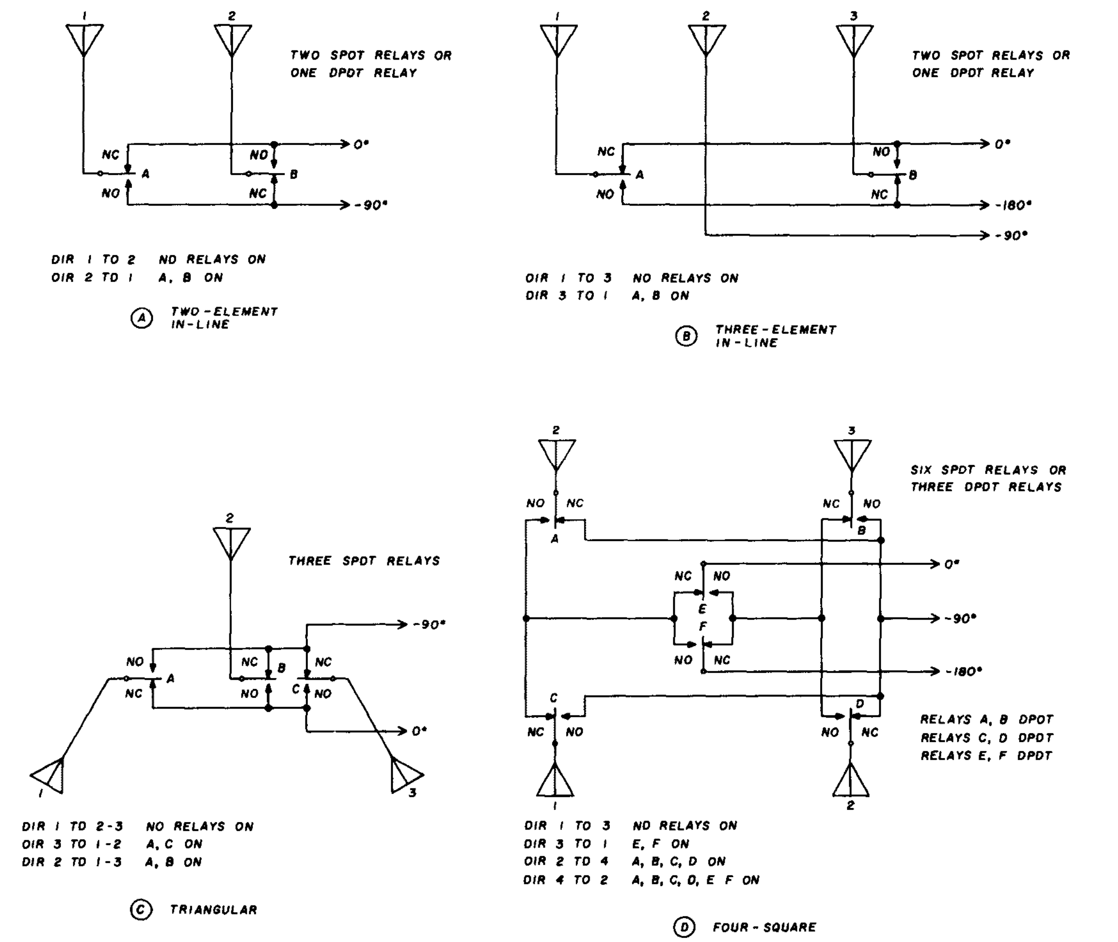


fig. 3. Relay interconnection diagram for four different vertical antenna arrays: (A) two-element in-line; (B) three-element in-line; (C) triangular configuration and (D) "4-square" configuration.

do not connect this chain to the shack line coax terminal until completing the adjustment of all chains.

In some circumstances it may be easier to simulate the calculated termination impedance of a network, in which case the bridge will look into this network in the same direction as the transmitter would. The impedance bridge is set to the pure resistance expected at the network input and it is connected to the shack line terminal. The simulated impedance (the same as calculated, not the *conjugate*) is connected to the coax terminal where the element feeder is normally connected. Assuming we are tuning the same network chain as above, a temporary connection of the shack line coax terminal is made to the input of the L-match (the same point at which the resistor was connected in the previous case). Except for these differences, the tuning procedure is the same. After tuning the L-match, the temporary bridge connection is transferred to the input of the Pi circuit, and an interconnection is made between the two networks in readiness for tuning the Pi circuit. After completion of tuning the Pi circuit, remove the shack line connection to it in preparation for tuning the next chain. Incidentally, there is nothing wrong with connecting the feeder coax into the chain to check out the entire network chain, making the appropriate changes to the bridge settings and/or the simulated network loads. However, do not connect an actual array element to this feeder in the expectation the element will present its array drive-point impedance, saving you the bother of simulating it. This simply won't work; part 4 of this series (October, 1983) explains why it will not.

directional switches

Relay interconnection diagrams for directionally switching four different vertical arrays is provided by **fig. 3**. When selecting relays for this application, remember that no one relay is switching *all* of your

transmitter power; consequently ratings may be safely reduced. Since RF is being handled, ceramic insulation is advised, though I found no problem with linen bakelite at 80 meters. Always avoid "hot" switching relays. Even if they can stand it, your linear will not — and neither will the network, since high voltages will be present during switching.

Photo 4 shows a 4-square array relay construction that use three small telephone-type DPDT military surplus relays. At first I lost several relays each summer due to sympathetic discharges from lightning which would burn out the solenoids. This was cured by connecting a silicon high current diode in reversed direction across the coil in parallel with a 0.1 μF ceramic disc capacitor. I have since lost a few diodes to these discharges, but no more relays. Failed diodes are "shorts," so the 28 VDC supply to this system requires a protective series resistance to guard against this possiblity and to prevent damage to the power supply from the discharge.

on rounding-off calculations

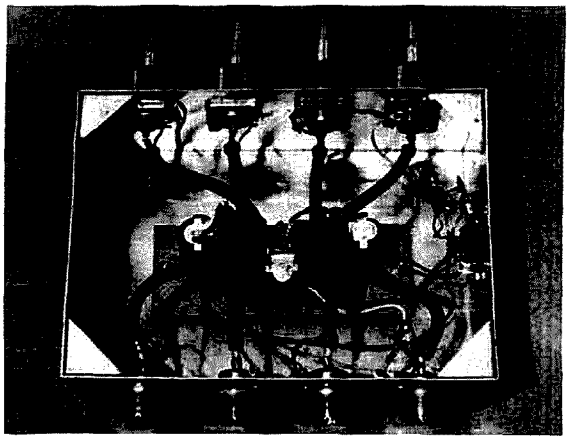
Calculator algorithms and computer operating system programs use guard digits as a means of retricting degradation of accuracy due to round-off in repetitive calculations. All values begin with and are calculated to one or two digits more than shown to the user. For example, if a calculator displays ten significant digits, it actually keeps values to eleven or twelve digits in its internal registers. Round-off errors thus tend to be restricted to these extra figures and seldom affect anything more than the least significant displayed digit. Since these extra digits rarely convey additional accuracy, they aren't displayed.

This concept also applies to calculations done by hand or with a slide rule. For example, using 3.14 for Pi reduces accuracy to only three significant figures *before* any calculations are done. A few computations immediately reduce that accuracy. Let's watch what happens with a simple calculation:

$(\text{Pi} \times 5)^2 - (\text{Pi} \times 78)$
$= 1.57$ if 3.14 is substituted for Pi.
If 3.1416 is used the result is 1.6965.
If 3.141593 then the result is 1.69591.

Note that the first approximation for Pi, accurate to three significant figures, has produced results accurate to only a *single* significant figure in just a few computations. What is happening is that every time this rounded off value for Pi is used, a small error is reintroduced, and in effect, the error is compounded.

Although many calculators round off to the nearest decimal, most small computers, and many large ones too, merely truncate values to some number of digits without adjustment to the nearest decimal, causing even more rapid divergence from accuracy if truncation is occurring after only a few digits.



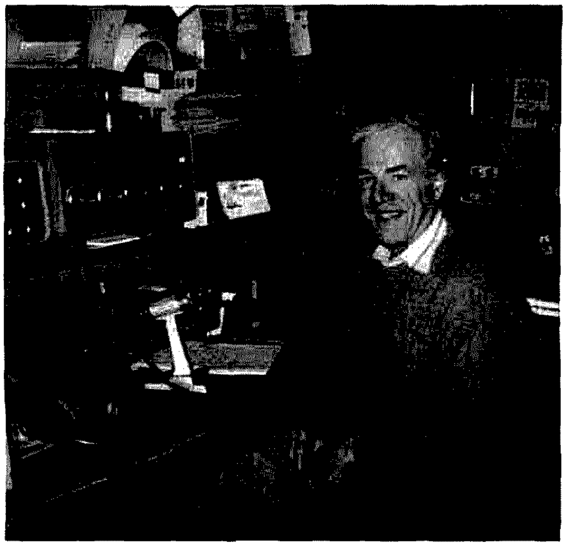
3 × 5 × 7 inch (7.6 × 12.7 × 17.8 cm) box houses all relays for switching four square vertical array.

The point of this discussion is to convince you to keep computations to several significant digits more than the accuracy you'd like to end up with. Except in determining rough approximations, constants such as 3.14 for Pi, or the number 984, expressing the speed of light in millions of feet per second, should always have two or more additional significant digits. This becomes particularly important when trigonometric functions involving angles approaching quadrant boundaries (for example, 0, 90, 180, or 270 degrees) are being used. Anyhow, in this day of ubiquitous calculators and computers, calculation to ten significant figures represents no personal mental effort.

The following constants, important to many calculations, are given to a level of precision more than sufficient for most applications:

Pi	3.141592654
e	2.718281828 Naperian logarithm base
c	299.792456×10^6 velocity of light, meters/second 983.571049×10^6 velocity of light, feet/second
cm/in	2.54 (exact) Metric to English unit conversion

Readers have inquired about the values given in these articles for inductances and capacitors in 4-terminal networks. Where, for instance, is a capacitor of value 734.8 pF to be obtained? Obviously no capacitor of that value will be listed in any catalogue; neither could we hope to find it to such accuracy by a measurement and selection process without also controlling temperature, humidity, aging, and so on. Measuring a capacitor or an inductance to just *three* places requires careful technique. I showed values to greater precision in an effort to prevent (mostly unsuccessfully) confusion caused by round-off errors if readers attempted to work backwards from my values for a capacitance or inductance to compute reactances, voltages and currents.



The author, beside his modest station, has proven that over 260 countries can be worked on 75 meters with legal power and a good antenna system.

concluding comments

I thoroughly enjoyed putting this series together, even though it required far more time and effort than I could possibly have imagined. I hope it proves useful and educational, though I'm not sure whether the author or his readers gained more! I tried to leaven the theoretical with the practical, well aware of the difficulties and pitfalls of doing so in such a technical subject.

I introduced the topic of matrix algebra as a tool of nearly limitless versatility that literally begs to be used. It not only reduces the tedium of network design calculations and simplifies transformation of one network to another, but also makes child's play out of the calculations of input/output conditions when networks are cascaded. It is particularly well suited to computer programmed calculations because the fundamental algorithms are unchanging; only the specific network parameter calculations differ. I did not begin to plumb the possibilities in these articles; there are ABCD parameters specific to lattices (bridge circuits), bridged Tee's, all types of transformers, real coax (with loss) and on and on. I sincerely hope this alone has found fertile imaginations in which to take root. It mystifies me that so powerful a tool has found so little welcome in our engineering educational institutions.

Antenna experimentation has always been of absorbing interest to Radio Amateurs, whether for DXing, for propagation studies, superior contesting, or to satisfy one's curiosity. Even though we have the ability, most of us don't have the time to devote to exploring the complexities of our station equipment. But anyone can innovate with a piece of wire, and would that it will always be so. However, antenna experimentation isn't magic; it's a technology like any other. Most of the fundamental principles were established two generations ago, though many of the pioneers, whom we find referenced and footnoted in articles and texts, are still with us.

To Bob Booth, WB6SXV, and Mason Logan, K4MT, for their encouragement, advice, and careful proofreading — which more than once kept me honest — much deserved words of appreciation. Finally, I want to thank you, my readers, for the many kind comments you sent my way via letters and on the air, and most of all — for your patient attention.

references

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ham radio

✓ 128

ham radio TECHNIQUES

Bill Orr W6SAI

Fifty years ago this month a young junior high school lad walked into the New York office of the newly-formed Federal Communications Commission. The nameplate on the door in the old Federal Building still said "Federal Radio Commission." The Commission room was a typical 1930's government office with walls of varnished wood, dusty, oversize furniture and a heating system going full-blast.

It was close to zero outside and there was still some snow on the window ledges. After filling out an application form and enduring an interminable wait, I was instructed to sit at a table next to a window and to prepare myself for the code test for the Amateur examination.

I was a bundle of nerves and the room was stifling hot. When the Radio Inspector wasn't looking I quietly opened the window for a breath of fresh air.

The exam began. First I sent a minute or two of Morse on the hand key mounted on the desk. No problem. My pulse was still racing as the Inspector handed out earphones, paper, and pencil to me and a few other nervous wretches aspiring to become Radio Amateurs. We were cautioned to listen to the code for a short time; the inspector would tell us when to begin copying.

It seemed easy enough. Code speed was 10 WPM and I had practiced for hours, sending and receiving, at 12 and 15 WPM. I started to write down the code coming at me out of the headphones.

Suddenly there was a loud racket and a flurry of wings as two of New York City's infamous pigeons landed on the windowsill, only a few inches away from my left arm. I shut my mind to the distraction and grimly forged ahead, copying the torrent of code issuing from the spring-wound code machine at the front of the room.

But the pigeons would have none of this. One of them boldly stepped over the window ledge and onto my arm and started pulling at the threads in my sweater. I gestured furiously with my free arm, trying to scare the bird away. But no. The second bird observed the commotion, jumped over the window ledge, and started walking about on my desk, finally coming to rest upon the paper I was trying to write on.

I pushed back my chair and beat away at the birds, who responded immediately. One flew out the window, squawking loudly; the other deposited a large lump of foul-smelling debris on my exam paper. I heard snickers from the other examinees around me, but I frantically scribbled on, no longer aware of where I was, or what I was copying. It seemed an inglorious end to all of my months of hard work!

After what felt like *centuries* passed, the code mercifully stopped. The Inspector, who had observed my battle with the birds, took my paper with a grin, as the second bird strutted about on the windowsill. I closed the window with a sigh.

Now the Inspector was grading the code test; in just a few moments I would know the results. Luck was with me — I had passed! The Inspector congratulated me and predicted that I would become a good CW man, able to copy successfully through the toughest QRM.

Fifty years have passed since that eventful day, which was not only a turning point in my life, but an introduction to the world of electronics and radio communication as well. If I had failed the exam, would I have found the fortitude to take it again? Would fate have turned me toward a different profession?

On that cold spring day in 1934, two pigeons came very close to irrevocably altering my destiny. Today — a half century *later — I still don't like* pigeons.

new sources of RFI

The VCR is a popular item, to be sure, and more and more of them are appearing in American homes. But along with the VCR comes interference caused by the intrusion of HF and VHF signals into the recording circuits: complaints range from blanking out of the picture to herringbone pattern interference and an absence of color. In the majority of cases, the interference was traced to a nearby Amateur operating on the 80-meter band.

The problem is complex. Modern VCRs are built on plastic frames enclosed in unshielded plastic boxes, leaving them wide open to RFI intrusion (See Vaughn Martin's "EMI/RFI Shielding: New Techniques," Parts 1 and 2, *ham radio,* January and February, 1984). In addition, the black and white portion of the video signal is recorded onto the tape as an FM signal over a frequency range of 3.4 MHz to 4.4 MHz. 3.4 MHz is "sync-tip" and 4.4 MHz is "peak white." Thus the 80-meter band falls between black level and about fifty percent white.

According to Bill Bowen, K8YGT, of the JVC Company of America, the

only sure cure for the problem is proper shielding of the video head drum and head amplifier, and the addition of a line filter to the VCR. Even with all these precautions, some of the newer all-plastic chassis VCRs are almost impossible to clean up!

As an example, Bill cites his work on cleaning up an RCA VCT400, a 1979 model made for RCA by Panasonic. After some effort, Bill obtained a shielding kit from RCA that, when installed, solved the problem. It is suggested, therefore, that unless you are an expert on the innards of a VCR, the prudent thing to do is to contact the supplier and find out if a filter kit is available.

Bill says that a quick way to get a "fix" on a VCR's resistance to RFI is to take a 2-meter HT into the store and transmit while the VCR is running. Although the frequency of the HT is far removed from the sensitive 80-meter area, when operated closely to the VCR it may produce RFI on the tape. This should be helpful in selecting a model that would be relatively "clean" of interference.

I should add that in any case, a good line filter on the VCR should help because RFI travels the power lines from ham transmitter to TV or VCR. If you have a stereo, TV, or VCR in your house, a line filter on your transmitter (and possibly on the entertainment equipment as well) is mandatory for peaceful coexistence with your family.

video disc players

The CED-type of video disc players (such as the RCA models) can be affected by even very low levels of RF in the 900-925 MHz region. In fact, one store that displayed a video disc player couldn't get it to play at all. The problem was finally solved by a technician who discovered that the "Sens-o-Matic" anti-shoplifting system was causing the interference. This device consists of an RF proximity alarm that sounds whenever merchandise tagged with a small series-tuned trap disturbs the RF field at a sensor, usually located near a store exit. In the case of the disc player, the 908 MHz carrier radiated by the sensor was invading the disc player through the stylus flyback circuit. Retuning the disc player circuit solved the problem.

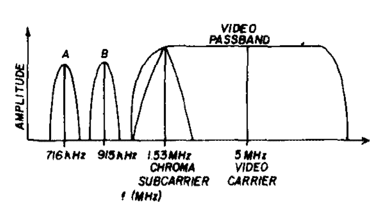

fig. 1. Frequency response of disc player covers range from audio carriers (A and B) at 716 and 915 MHz to video passband centered at 5 MHz. System is sensitive to 4 MHz as well as 910 MHz signals.

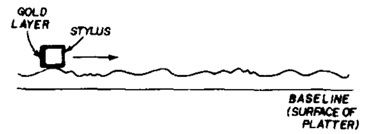

fig. 2. Cross-section (lengthwise) of video disc groove showing position of stylus and baseline. As the stylus moves up and down during playback, a varying capacitance is created between the gold layer at the back of the stylus and the platter. The capacitor tunes a 910 MHz circuit in the player.

how the disc player works

When a disc is cut, a combined video/audio signal on an FM subcarrier modulates a piezoelectric cutter, causing the depth of the grooves to vary with respect to the baseline (**fig. 1**). The audio carriers are centered about 716 kHz and 915 kHz, the video carrier is centered about 5 MHz, and the chroma subcarrier is at 1.53 MHz.

A lengthwise view of the platter groove is shown in **fig. 2.** During playback, as the stylus moves up and down, a varying capacitance is created between the gold layer at the back of the stylus and the player. This capacitance is one element of a 910 MHz series-tuned circuit that works somewhat on the principle of slope detection of an FM signal. The recovered signal is then split into its components (audio, video, chroma) for processing.

The sensitive portion of the machine is the flylead that connects the stylus to the circuit, which is a large portion of a wavelength long. Even though the whole circuit is enclosed in a two-compartment metal box and seems well shielded, the flyhead is very sensitive to RF in the 900 MHz region.

There will soon be a lot of RF activity — Amateur and otherwise — in the 900 MHz region. Cases of video disc RFI are already appearing in homes near military installations where 900 MHz radar is operating.

So far, there seems to be no easy solution to the problem. The basic circuit of the playback mechanism is shown in **fig. 3.** Any ideas from readers on this interesting and vexing problem would be appreciated.

light bulb RFI

What's next? A recent article in a broadcast industry newspaper pointed out that "the potential of RF lighting devices to cause interference to AM radio and certain other broadcast services has given rise to a regulatory debate at the FCC.

"Most broadcasters want the Commission to impose federal regulations on RF lighting devices, according to comments filed in response to an FCC Notice of Inquiry. But the FCC and most lighting producers would prefer to see the interference potential of these lighting devices controlled by a voluntary industry standard."

What are these lighting devices? I understand they are oversize bulbs that are not incandescent lamps, but are instead a variety of fluorescent lamp having a standard socket and a folded tube that is excited by a built-in VLF oscillator. In some versions of this RF-excited lamp, the oscillator runs continuously when the lamp is on; in others, the oscillator stops when the lamp is lit.

In either case, according to the article, "tests at the Commission show RF

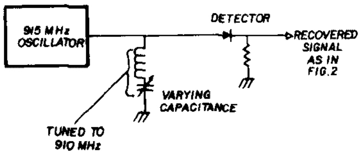

fig. 3. As the stylus moves up and down, a varying capacitance created between the stylus and the platter tunes a 910 MHz circuit, acting on the same principle as the slope detector of an FM signal. The recovered signal is then split into its component parts for processing (audio, video and chroma). Signals leaking into this circuit can upset the process and alter the video picture.

lighting emissions to be very broadband, with measurable emanations ranging from 10 kHz to 80 MHz.

"Referring to tests conducted by Harold Kassens of A.D. Ring and Associates, the NRBA found the interference potential of RF lighting 'alarming.' According to preliminary reports, AM radios were able to pick up objectionable radiation interference from as near as several meters from RF light bulbs.

"Mr. Kassens warns that if the development of this new technology is not regulated, it is the public which will suffer most when the uncontrolled use of the new product interferes with the reception of established broadcast services".

The article continues, "While broadcasters were skeptical about the lighting industry's willingness to establish an industry standard, General Electric, a leading manufacturer of RF lighting devices, expressed enthusiasm over voluntary limits." General Electric, moreover, presented proposed limits on radiation and measurement techniques for use in determining whether RF lighting devices meet the proposed limits.

It seems to me that Radio Amateurs would be vitally interested in this new RFI hazard. The Docket number on the FCC Notice of Inquiry is GEN 83-806; a copy can be obtained by calling the FCC at (202) 653-8247.

ham radio

easy antenna matching

Use a home computer and a noise bridge to solve for T-network values

Matching a transmission line to an unknown impedance can be done either by trial and error (probably most common method) or by measuring the actual impedance and designing a matching network. The latter method is faster and the availability of low cost noise bridges makes it much easier.

One of these noise bridges is the MFJ 202B. It comes with a very explicit manual and is easy to use except for one thing: the equations for converting from indicated values to actual impedances are a bit unwieldly. I took one look at those and decided that there was no way I'd do *that* more than once; instead, I wrote a program to do the calculations by computer. From there, it made sense to have the computer calculate the matching network, provide a printout of several measurements, and graphically summarize the data.

The program shown in **fig. 1** does all these things. It is written for the Radio Shack Color Computer™* but will run (except for the color graphics) as is on the TRS-80 Model 1™* and, with minor modifications, on any of the other popular machines. Its self-prompting feature makes it easy to use.

using the program

First, make the actual measurements. Write down, in order of increasing frequency, the frequency, indicated resistance, and indicated reactance. Treat capacitive reactance as a negative value.

Then load and RUN the program. The program will ask for the line impedance (used for SWR calculations) and then for frequency in MHz. Next you will be asked for the indicated values of resistance and reactance.

The screen will display the calculated impedance and SWR. (These values are also being stored internally for later use.) You will then have the opportunity to input more data, obtain a hard copy, calculate a matching network, or see a graphic display.

To input more data, merely repeat the process above. To calculate a matching network, answer "YES" when the question "MATCH?" appears on the screen. The display will now show the reactances to be used in the network of **fig. 2** and the necessary *component* values. You *can* influence the bandwidth and component values by specifying the value of Q to be used in the network. If the reactance for L1 is negative, a capacitor will have to be used.

To obtain a printout of the impedances as a function of frequency, answer "YES" to the "PRINT?" prompt. The printer first gives you the opportunity to name the antenna and then proceeds to tabulate.

To obtain a color graph of impedance versus frequency, answer "YES" to the "PLOT?" prompt. The screen will clear, axes will be drawn, and real and reactive impedances will be plotted. The resistance plot will consist of yellow blocks; the reactance plot, white. When the display is complete, the program will stop and wait for you to press any key. This allows you time to evaluate the plot. Pressing any key will resume program operation and erase the graph. At the completion of any of these operations, you will again be given the opportunity to input more data, calculate a match, and so forth.

To keep the program simple there is not a large

*Radio Shack Color Computer and TRS-80 Model 1 are trademarks of the Tandy Corporation.

By James A. Sanford, WB4GCS, 248 Worden Street, Portsmouth, Rhode Island 02840.

amount of error-trapping, so your data must be reasonable. For example, the program assumes the antenna impedance is *lower* than the line impedance — this is certainly true of most antennas I've needed to measure. The matching network calculation does include any antenna reactance; if there is no reactance at all, there is not much point in plotting the response: the plot will not work if there is no reactive data. If the data is not entered in order of increasing frequency, the printer output will not be in the proper order.

```
1 DIMF(20),R(20),X(20)
     :K=1
5 PI=3.1415926535
10 CLS
     :PRINT@224,"MFJ NOISE BRIDGE CALCULATOR"
     :PRINT"BY JAMES A. SANFORO"
     :PRINT
     :PRINT
     :PRINT
     :INPUT"LINE IMPEDANCE (OHMS)";ZL
15 FOR I = 1 TO 500
     :NEXTI
20 CLS
     :INPUT"FREQUENCY (MHZ)";F
     :INPUT"RESISTANCE (OHMS)";RD
     :INPUT"REACTANCE (PF, XC NEGATIVE)";XD
     :INPUT "EXPANDER";E$
     :PRINT
     :PRINT
     :PRINT
     :IF E$="YES"THENGOTO1000
30 XU=(888/F)-(160000/(F*(180+XD)))
     :RU=RD
39 S=SQR(RU^2+XU^2)/ZL
     :IFS<1THENS=1/S
40 PRINT"IMPEDANCE = ";RU;" + ";INT(100*XU)/100;"J OHMS";
     :PRINTTAB(10);"= ";INT(100*SQR(RU^2+XU^2))/100;" ANGLE ";INT(100*(ATN(XU/RU
))*(180/3.1415926535))/100
     :PRINT"SWR = ";S
45 S(K)=S
     :F(K)=F
     :R(K)=RU
     :X(K)=XU
     :K=K+1
47 INPUT"ANOTHER RUN";R$
     :IFR$="YES"THEN20ELSEINPUT"PRINT";R$
     :IFR$="YES"THEN2000ELSEINPUT"MATCH";R$
     :IFR$="YES"THEN3000ELSEINPUT"PLOT";R$
     :IFR$="YES"THEN4000ELSECLS
     :END
1000 XEQ=888*XD/(F*(XD+180))
     :XU=(40000*XEQ/((200-RD)^2+XEQ^2))
     :RU=200*(200*RD-RD^2-XEQ^2)/((200-RD)^2+XEQ^2)
     :GOTO39
2000 INPUT"ANTENNA";A$
     :PRINT#-2,TAB((80-(LEN(A$)))/2);A$
     :PRINT#-2,CHR$(10);CHR$(10),CHR$(10);"NUMBER","FREQUENCY","REAL","IMAGINARY
","SWR";CHR$(10)
2010 FORJ=1TOK-1
     :PRINT#-2,TAB(2);J,F(J),R(J),X(J),S(J)
     :NEXTJ
     :PRINT#-2,CHR$(10);CHR$(10);"FEEDLINE IMPEDANCE = ";ZL;" OHMS"
     :PRINT#-2,CHR$(&H0C)
     :GOTO47
3000 CLS
     :INPUT"CHOOSE Q FACTOR";Q
     :A=RU*(1+Q^2)
     :B=SQR(A/ZL-1)
     :X1=RU*Q
```

fig. 1. Antenna matching program is written for Radio Shack Color Computer™ but will run on TRS-80 Model 1™ or, with modification, on any other personal computer.

Some other comments regarding interpretation of the data are appropriate. First, the MFJ 202B is not intended to be a laboratory instrument. Even though the computer prints out answers to 9 digits, there is no point in believing anything past the first or second digit after the decimal point. This is simply because the input data is not that good.

Second is the issue of SWR. Much has been written about misconceptions of SWR; suffice to say that SWR is an *indication* of the impedance match. Some

```
      :X2=ZL*B
      :XC=A/(Q+B)
3010 PRINT"REACTANCES OF
      :"
      :PRINTTAB(5);"L1 = ";X1-XU;" OHMS"
      :PRINTTAB(5);"L2 = ";X2;" OHMS"
      :PRINTTAB(5);" C = ";XC;" OHMS"
3015 PRINT"VALUES OF
      :"
      :PRINTTAB(5);"L1 = ";
      :IF X1+XU>0THENPRINT (X1+XU)/(2*PI*F);"MICROHENRIES"ELSEPRINT1E6*1/(2*PI*F*
(XU+X1));"PICOFARADS"
3016 PRINTTAB(5);"L2 = ";X2/(2*PI*F);" MICROHENRIES"
3017 PRINTTAB(5);" C = ";1E6*1/(2*PI*F*XC);"PICOFARADS
3020 GOTO47
4000 CLS0
      :FOR I = 1 TO 417 STEP 32
      :PRINT @I,"! ";
      :NEXTI
      :PRINT@448,STRING$(31,"-")
      :L=64
      :FOR Q = 1 TO9
      :READ D$
      :PRINT@L,D$;
      :L=L+32
      :NEXT
      :DATA I,M,P,E,D,A,N,C,E
4010 PRINT@490,"FREQUENCY";
4020 BW=F(K-1)-F(1)
      :FOR I = 1 TO K-1
4021 IF RH<R(I) THEN RH = R(I)
4022 IF RL>R(I) THEN RL = R(I)
4023 NEXTI
4025 FOR I = 1 TO K-1
4026 IF X(I)>P THEN P = X(I)
      :
4027 IF X(I)<T THEN T =X(I)
4028 NEXTI
4030 PRINT@480,F(1);
      :PRINT@505,F(K-1);
      :PRINT@32,INT(RH);
      :PRINT@57,INT(P);
      :PRINT@416,INT(RL);
      :PRINT@443,INT(T);
4035 FORI=1TOK-1
      :W=2+(F(I)-F(1))*(61/BW)
      :Y=28-28*INT(R(I)-RL)/(RH-RL)
      :SET(W,Y,2)
      :Y=28-28*INT(X(I)-T)/(P-T)
      :SET(W,Y,5)
      :NEXTI
4037 PRINT@13,"R";
      :PRINT@17,"X";
      :SET(32,0,2)
      :SET(40,0,5)
4040 IFINKEY$=""THEN4040ELSECLS
      :GOTO47
5000 '************************ THIS PROGRAM CALCULATES RESULTS USING READINGS
FROM MFJ NOISE  BRIDGE, MATCHING NETWORKS, AND  SHOWS A GRAPH OF RESPONSE.
      <RUN> IT TO USE.
```

fig. 1. (continued)

John Bean

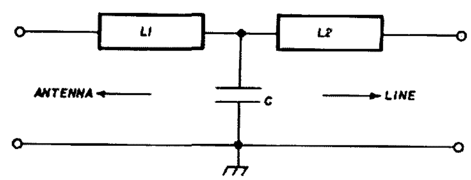

fig. 2. T-network used in this case to match antenna impedances lower than 50 ohms.

transmitters may not tolerate the reactive load even though the SWR is not excessive.

Third, temper results with reality. If a matching network calls for 10 Henries and 0.0001 picofarads, try a different value of Q, or change the effective line impedance with a broadband transformer.

Although I have talked about measuring antenna impedances, the MFJ bridge and these techniques can be used to measure other parameters. Linear amplifier input impedance can be measured and even the input impedance of a properly biased transistor stage can be determined. This could make matching easier. Realize that the MFJ bridge has limits on what it can measure, and that large voltage or RF will destroy it. So don't try to measure the plate impedance of a 2-kilowatt linear at 4 GHz!

how it works

Program operation is straightforward. Lines 1 through 10 initialize the program. Line 20 allows input of the measured information. If the noise bridge expanded range is not used, line 30 calculates the actual impedances. Otherwise, this is done by line 1000. These calculations are in accordance with the equations in the MFJ manual (see appendix). Line 39 calculates the magnitude of the SWR. Line 40 displays the result and line 45 updates the stored data. Line 47 gives you the opportunity to select what to do next. The printer output routine is at line 2000 and the matching network calculations at line 3000. The equations used are shown in the appendix.

The plotting routine begins at line 4000. For use in the TRS-80 Model I or III, simply change the scaling calculations and delete references to colors. To plot on other color computers, check the User's Manual that accompanies your machine.

conclusion

Impedance matching can be simplified by using a noise bridge to make actual measurements and a computer to do the brute force number-crunching. This program will help in that effort. Now, matching strange new antennas can be easier and a lot less frustrating.

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3. *Reference Data for Radio Engineers*, 6th Edition, Howard W. Sams Co., Indianapolis, Indiana, 1979.
4. J. J. DeFrance, *Communications Electronics Circuits*, Rinehart Press, 5643 Paradise Drive, Corte Madera, California 94925, 1972.

appendix

1. Calculating reactance without expander:

$$R = Rd$$
$$X = \frac{888}{f} - \frac{160{,}000}{f(180 + Cd)}$$

where f = frequency in MHz
Cd = reactance dial reading in pF
− for capacitance, + for inductance
X = reactance in ohms
R = resistance in ohms
Rd = indicated resistance

2. Calculating reactance with expander:

$$Xeq = \frac{(888)\,(Cd)}{f(Cd + 180)}$$
$$R = \frac{200\,[200\,Rd - Rd^2 - (Xeq)^2]}{(200 - Rd)^2 + (Xeq)^2}$$
$$X = \frac{40{,}000\,Xeq}{(200 - Rd)^2 + (Xeq)^2}$$

3. Three element matching network:

$$X_{L1} = R_A\,Q$$
$$X_{L2} = Z_{Line}\sqrt{\frac{R_A\,(1 + Q^2)}{Z_{Line}} - 1}$$
$$X_C = \frac{R_A(1 + Q^2)}{Q + \sqrt{\frac{R_A\,(1 + Q^2)}{Z_{Line}} - 1}}$$

where R_A = antenna resistance
Z_L = line impedance
Q = quality factor
X = reactance

For a copy of the program on tape, send $10.00 to James A. Sanford, 248 Worden Street, Portsmouth, Rhode Island 02840.

ham radio

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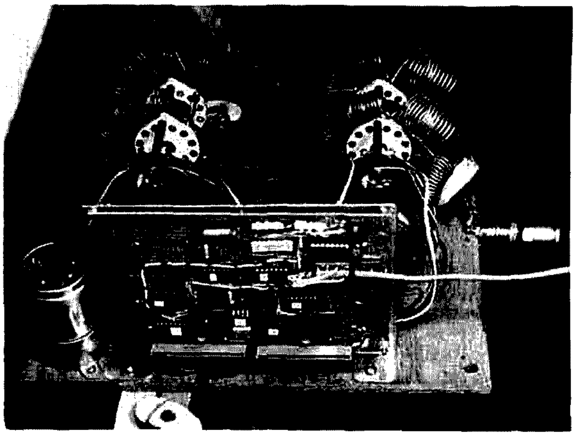

an end-fed multiband 8JK

A new approach to tuning and feeding for multiband operation

A wire that is 2-1/4 wavelengths long at the high-frequency end of the 10 meter band can be provided with adjustable series inductive loading to give a low feed impedance throughout the 10, 15, and 20 meter bands. Two such wires in an "end-fed W8JK" arrangement can form a multiband antenna whose dimensions and feed arrangements are well-suited to small backyards, and which combines useful gain with an SWR better than 1.4:1 across the whole of the three bands. Further possibilities exist for using the basic element and feed arrangement in other configurations.

design requirements

Several years of experimentation and experience convinced me that an HF band antenna for my location had to meet five requirements.

First, since I live in the suburbs, it had to be a fixed wire with unobtrusive supports, optimized to perform at a maximum height of about 30 feet.

Second, it had to be multiband so that I could follow my changing interests and also respond to varying propagation conditions. However, since rotatable arrays have already been ruled out by the first requirement, and I have no specialized geographical interests, there is no particular merit in having the same horizontal polar diagram on each band. Therefore, "long wire" arrangements were quite acceptable — particularly if they avoided the losses inherent in trapped dipoles.

Third, end feed was desired. My "shack" is inside the pitched roof of my house, quite near to the end of any practical long wire. Feeder cables running across the lawn to a center feed point are unsightly and subject to serious losses in humid weather. (I used to think that a 300 ohm tubular feed was ideal until I emptied a pint of water out of a length of such a feeder that was tied along a fence!)

Fourth, the feed should be a balanced two-wire arrangement. Early experiments with a single wire fed against ground and with Zepp feed had produced RFI to both household electronics and to my equipment in the shack.

Fifth, a little gain would be nice! Rotatable arrays get gain by narrowing the horizontal beamwidth, but in the case of a fixed wire antenna intended to provide coverage of all directions, this is impossible. The main option is to shape the vertical radiation pattern so as to concentrate energy into useful angles of elevation, ideally between 10 and 20 degrees above the horizontal — and minimize ohmic losses in the antenna

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and in the surrounding ground. A second possibility is to use two or more antennas driven with different phases so as to provide steerable maxima and nulls. This is a choice that may be considered *after* the basic antenna is selected. Therefore, the choice was limited to arrangements that provide zero radiation both up and down — that is, the three families shown in **fig. 1**.

Study of the literature disclosed no clear evidence to support the choice of vertical polarization on the HF bands at the heights proposed, although, since the performance of horizontal antennas is more strongly affected by height, there might be a good case for a vertical antenna if the available height were much less. In a vertical antenna such as that shown in **fig. 1A**, the current is zero at the top and so the effective height is less than the actual height. If the radiator is trapped for multiband operation the top section may not be used, further reducing the effective height and with it the efficiency.

The so-called "Lazy H" array (**fig. 1B**) uses two dipoles stacked one above the other and fed in phase so as to provide maximum radiation horizontally and minimum radiation in the vertical direction. However, the effective height of the array is only the average height of the two elements, and the lower element would be inconveniently near ground level.

In contrast, the effective height of the "W8JK" array (**fig. 1C**) equals its actual height. The two dipoles are fed out of phase, and so straight-up-and-straight-down radiation is zero regardless of spacing, giving a potential for multiband operation with high efficiency. However, all the published multiband designs that I could find used center feed. So the question became, "how can two parallel long wires be end-fed out of phase?"

the horizontal loop

I spent a long time experimenting with the horizontal loop antenna of **fig. 2**. A 40-meter length of wire formed into a loop presents a low impedance when its length is any multiple of a full wavelength: in this case it is resonant at 7, 14, 21, and 28 MHz. Unfortunately, the resonances follow the simple harmonic relationship only when the wire forms a circular loop. For more complex shapes, the resulting concentrations of inductance and capacitance disturb the relative resonant frequencies so much that an ATU is still needed for multiband operation. Furthermore, while the currents in opposite sides of the loop flow in opposite directions (as is required for "W8JK" operation) it is not possible to find a shape where the whole length of the opposite sides are spaced by the 0.1 wavelength to 0.25 wavelength that is desirable. For example, in the rectangular loop of **fig. 2**, although the *long* sides are spaced correctly there are also current maxima in the *short* sides, which are spaced by 1.5 wavelength at

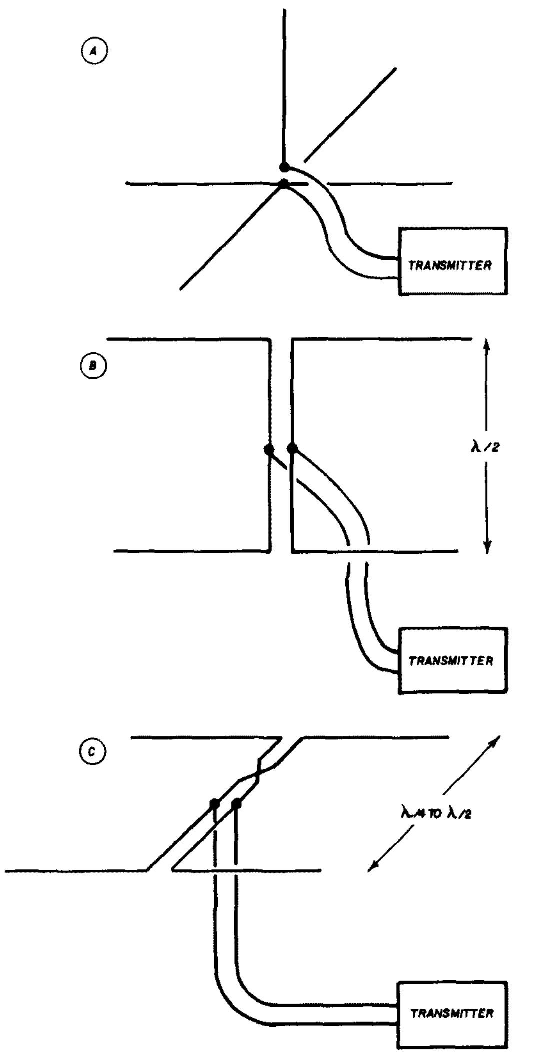


fig. 1. Basic antennas that don't radiate straight up or down: (A) vertical 1/4 or 5/8 wavelength with radials or groundplane; (B) horizontal dipoles ("lazy-H"), vertically stacked fed in phase; (C) horizontal dipoles (W8JK) horizontally stacked fed out of phase.

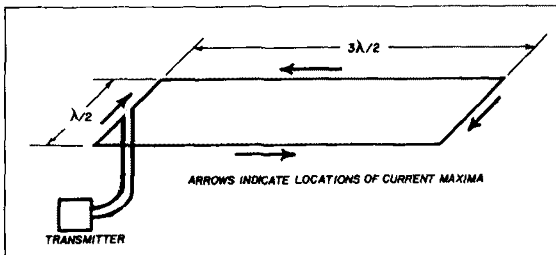


fig. 2. The 40-meter circumference horizontal loop antenna.

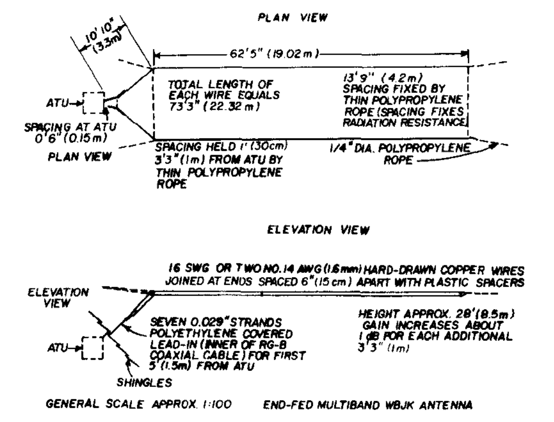


fig. 3. The end-fed multiband W8JK antenna.

30 MHz, and so provide strong horizontally-polarized end-fire radiation at directions 70 degrees above and below the horizontal plane, with a consequent loss of gain in useful directions.

the end-fed W8JK

At about this stage I read G6XN's book.[1] His words confirmed my faith in the W8JK type of antenna, gave me new insights about how to feed the end of an antenna, and provided useful cautions about the RF resistance of long wires. With confidence restored, I settled down to think about ways of eliminating the radiation from the short sides of the loop antenna, and came up with the arrangement of **fig. 3**. This retains both the basic horizontal radiation pattern of an end-fed long wire (2 wavelengths at 28 MHz: 1-1/2 wavelengths at 21 MHz and 1 wavelength at 14 MHz) and the gain of the W8JK, but has much reduced end-fire radiation. It will be seen that the far end of the loop has been removed altogether since it was not meeting its original objective of simplifying the tuning. The near-end "fan-out" from the ATU to the parallel wires has been shaped to minimize its end-fire radiation. The design of this section is based on study of the voltage distribution along either wire of the antenna, which is plotted in **fig. 4**. It can be seen that in every case the voltage is high, and hence the current is low, at a distance of about 19 to 21.5 meters from the open ends. It is in this region that the most rapid "fan out" of the feeder may be allowed without too much wasted high angle radiation.

antenna tuning unit

The length of each antenna wire is chosen so that at the highest design frequency (29.7 MHz), the antenna is resonant; that is the feed voltage is at a minimum. This occurs 22.32 meters from the open end; at this point the wires are terminated just inside the pitched roof of the house, at an antenna tuning unit that is in a dry location and easily accessible. In fact, at 29.7 MHz no tuning is required since the spacing of the parallel section of the antenna can be adjusted to obtain a radiation resistance of 50 ohms. At lower frequencies the effective length of the wires must be increased to obtain resonance; it can be seen from **fig. 4** that 1.5 meters must be added at 28 MHz, and about 2.4 meters and 4.3 meters, respectively, are required for 21 and 14 MHz operation. In practice this length can be added by the ATU in the form of adjustable series inductances. On the lower frequency bands radiation resistance decreases with smaller array size. On 14 MHz it was measured at about 25 ohms. However, the capacitance of the ATU and fan-out section can be chosen so that with the ATU inductance it forms a matching section that gives a low SWR on all three bands.

While a 2-gang roller inductor would be the ideal form of variable inductance to provide the balanced tuning network that is necessary, one was not available. Instead, "units" and "tens" steps of inductance were provided by two 2-gang ceramic wafer switches (driven by semi-rotary solenoids to be described shortly). As can be seen in the photograph (**fig. 5**) the coils

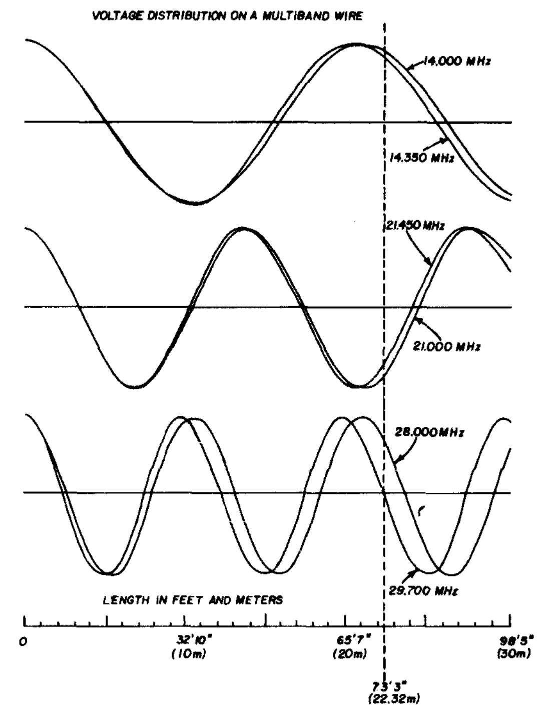
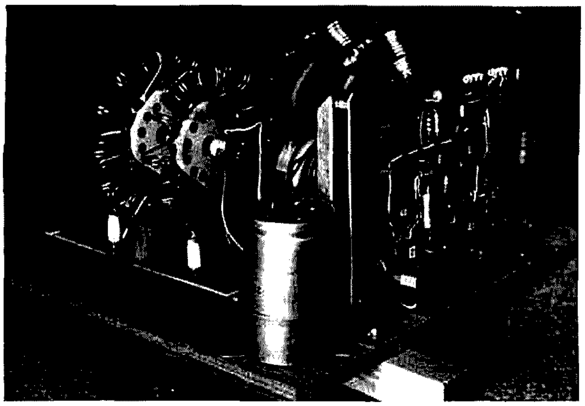


fig. 4. The voltage distribution along an end-fed long wire on the 20 meter, 15 meter, and 10 meter bands.



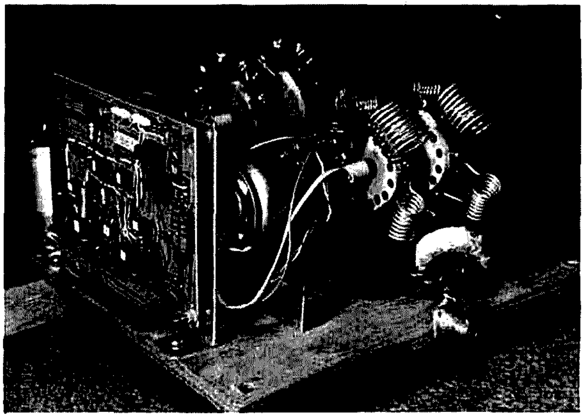


fig. 5. Side views of the remotely controlled antenna tuning unit.

are connected directly to the wafers, and the wafers spaced approximately 1.5 inches (40 mm) apart to optimize the stray capacitance. **Fig. 6** shows that an SWR of better than 1.4 can be obtained across almost all of the three bands by varying the inductance alone; no variable capacitor is necessary.

The RF circuits of the ATU are shown in **fig. 7**. Immediately next to the antenna connections are two lightning arrestor gaps, connected by a substantial conductor to a nearby copper water pipe. Next comes the "units" inductance providing up to 0.9 μH in nine switched steps. This is followed by the "tens" inductance, whose switch is wired so that unused inductor sections are short-circuited to avoid losses due to resonance with stray capacitance. The balun is of W1JR's original "supertoroid" design.[2] The criticisms of K4KJ do not apply to applications such as this where there is substantial capacitance to ground in the antenna itself.[3] The balun should be wound with 50 ohm cable to ensure that no reflection occurs at the connection to the main 50 ohm cable. This is the cable that leads from the ATU to the transmitter (which in my case, is also within the pitched roof some 33 feet or 10 meters away from the ATU).

The bandwidth of this antenna system is quite large on 10 meters, where little or no series inductance is used, and a single inductance setting may be used over a 200 kHz portion of the band. On 20 meters, however, it is necessary to change the inductance switch settings every 20 kHz or so. Control of the "tens" and "units" inductors is therefore provided by two thumbwheel switches at the transmitter. DC power and the multiplexed BCD thumbwheel settings are sent along an 8 conductor cable to the ATU and used to position the two inductor switches.

ATU control circuits

The solenoids that drive the ATU switches require considerable power for the short time that is needed

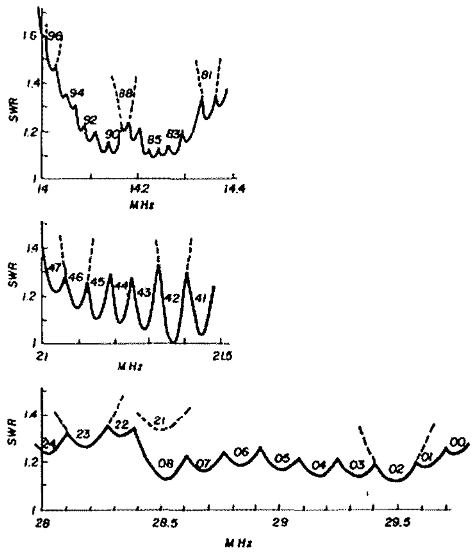


fig. 6. SWR plot achieved by the end-fed W8JK and series-inductance ATU. The numbers against each curve specify the 'tens' and 'units' inductance switch settings.

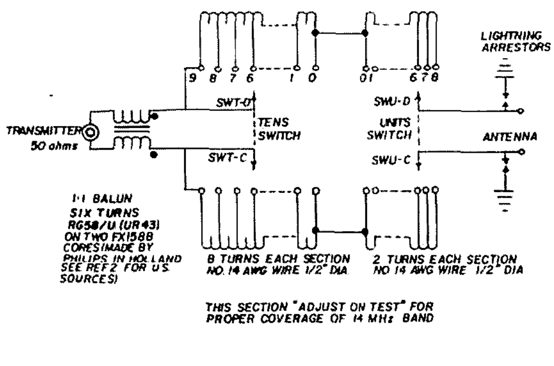
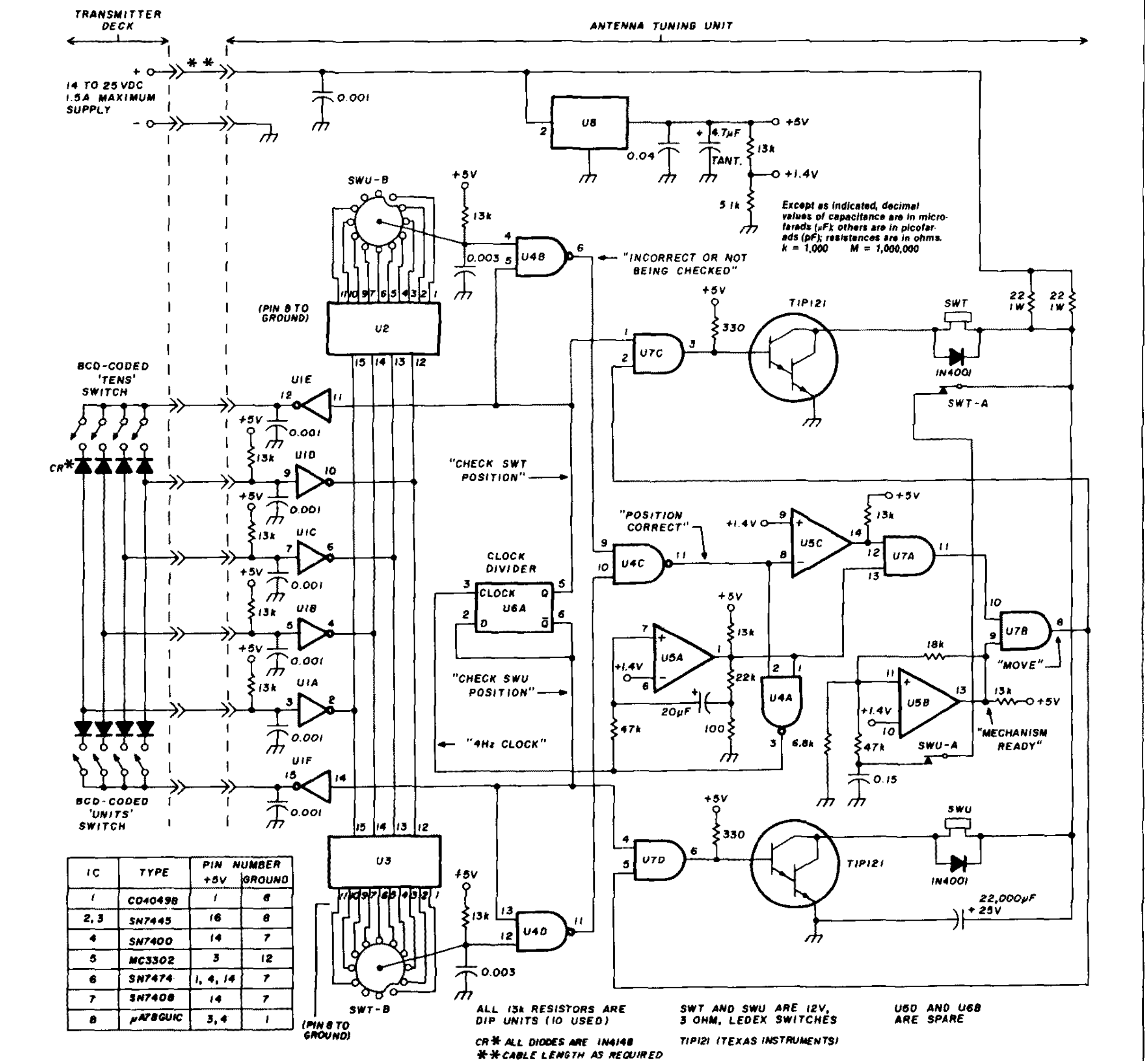


fig. 7. RF circuits of the series-inductance ATU.

to move the switch from one position to the next. The ones I used (SWT and SWU in **fig. 8**) need 4 amps at 12 volts, and so, to allow the use of thin telephone-type cable to the operating position, energy is stored at the ATU in a 22,000 microfarad capacitor. Circuit operation will be described with references to the "tens" switch SWT; operation of the "units" switch is similar but occurs on the other half-cycle of the clock waveform provided by divider U6A. When SWT is energized the solenoid pawl moves the switch ratchet to the next position and opens the interrupter contact SWT-A. In this circuit the interrupter contact tells the logic to turn off the power Darlington transistor that drives the solenoid; comparator U5B and its associated circuit then ensure that power cannot be reapplied until the pawl has returned to its rest position and the storage capacitor has been recharged to at least 14 volts.

The actual position of the "working" switch wafers SWT-C and SWT-D (**fig. 7**) is sensed by the all-but-one-connected wafer SWT-B (**fig. 8**) which is driven from the BCD — decimal converter U2. If the position is incorrect, then on the appropriate phase of the clock the output of U4C will be low. This stops the clock oscillator (U4A and U5A) and allows the solenoid to step repeatedly under the control of U5B until the correct position is reached. The clock oscillator then restarts, so that the position of the other sole-



IC	TYPE	PIN NUMBER +5V	PIN NUMBER GROUND
1	CD4049B	1	8
2, 3	SN7445	16	8
4	SN7400	14	7
5	MC3302	3	12
6	SN7474	1, 4, 14	7
7	SN7408	14	7
8	µA78GUIC	3, 4	1

fig. 8. Control circuits of the series-inductance ATU.

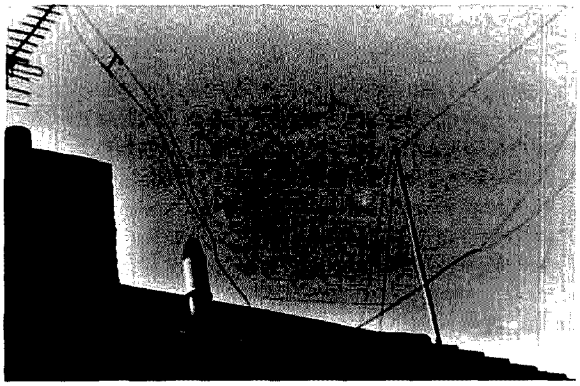


fig. 9. Feed-end of the antenna.

noid may be checked. To minimize interference that might be coupled into the antenna circuit, the clock oscillator is run at the lowest frequency that does not noticeably degrade the response of the circuit to changes of thumbwheel position — that is, 4 Hz. Consideration of the threshold levels resulting from the diode "or" circuit associated with the thumbwheel switches shows that a CMOS interface U1 must be used, even though the other circuits are all TTL. This is unfortunate, since the circuit is exposed to induced transients from lightning. It is recommended that a socket be used to mount the chip. The position-sensing switch wafers are not shielded from the high RF voltages on the working wafers; with 100 watts of RF it proved necessary to add the 3 nF capacitors to the rotors of SWT-B and SWU-B. Shielding is clearly desirable and would probably be essential with higher power.

choice of antenna conductor size

As previously mentioned, the radiation resistance at 14 MHz is only 25 ohms, and care is needed to ensure that resistive losses in the 44.64 meters of resonant radiating elements are not significant. The lowest frequency band represents the worst case, since measurement of this antenna has shown that the radiation resistance is roughly proportional to frequency, whereas the wire loss resistance due to skin effect only increases with the square root of frequency. With the two parallel No. 14 AWG (1.6 mm) conductors specified for each element (see **fig. 3**) the effective resistance is about 2.5 ohms giving less than 1 dB loss. The 1.5 meters of each element nearest the ATU are made from polyethylene-covered stranded conductors to provide good insulation and flexibility at the point of entry though the roof structure as shown in **fig. 9**.

conclusion

This antenna performs much like any other long wire; it would be difficult to make comparisons on the

✓ 188

Amateur bands with sufficient accuracy to make any other claim. However, its losses are small and its directivity well defined, and therefore theory would suggest 3 to 5 dB gain relative to a dipole in the best directions. Its real advantages are the balanced end feed and multiband operation.

However, the principle of the tuning unit may be applied to many other antennas. What has been described is a way of using a single balanced variable to tune and match two long wire elements. It has been suggested to me that if these wires formed an Echelon array then one of the elements could be moved to provide steerable nulls. A "V" beam is possible, although the angle would have to be a compromise between the three bands. With the two elements co-linear, the design should display characteristics similar to the G5RV and offer the advantage of an alternative feed arrangement.

Finally, it may be noted that the antenna as described in **fig. 3** is also exactly resonant at 10 MHz, though its radiation resistance is very low and a different tuning arrangement is necessary.

references

1. L.A. Moxon, G6XN, *HF Antennas for All Locations*, Radio Society of Great Britain, 1982.
2. Joseph Reisert, W1JR, "Broadband Balun," *ham radio*, September, 1978, pages 12-15.
3. J.J. Nagle, K4KJ, "High Performance Broadband Balun," *ham radio*, February, 1980, pages 28-34.

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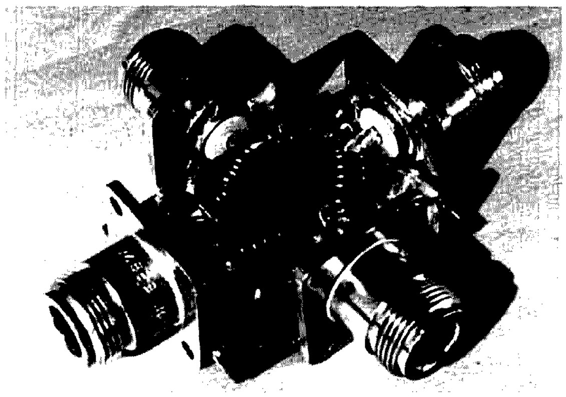
Applications include power splitting/combining, impedance transformation and attenuation

the branch-line hybrid: part 2



At frequencies below 30 MHz, lumped constants offer less loss than coaxial cable to form compact hybrids.

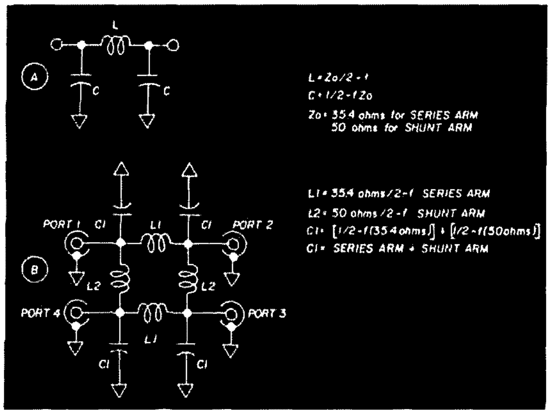


fig. 1. Lumped-constant branch-line coupler is formed with an artificial transmission line in series (A) and shunt (B) with each port.

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lumped-constant branch line

Below 100 MHz the coaxial cable form of a branch-line hybrid becomes excessively bulky. At these frequencies the branch-line hybrid can be formed using capacitors and inductors. An artificial transmission line may be formed using either the tee or π network. The π network is normally chosen because it functions as a lowpass filter. The inductance and capacitance values given in **fig. 1A** are calculated using the characteristic impedance of the transmission line. The single-section branch-line coupler (**fig. 1B**), is formed by two series and two shunt artificial transmission lines. Inductance L1 in the series line is calculated using the 35.4-ohm characteristic impedance calculated earlier. The 50-ohm shunt line is formed using L2. The capacitors, C, from each artificial line are added together to form a composite value, C1. The element values for a 3-dB 50 ohm branch-line hybrid are given in **table 1**. The theoretical and experimental response of the lumped-constant hybrid is shown in **fig. 2**. Note that the bandwidth is only slightly reduced from that in the transmission-line case. The curves are slightly skewed when using lumped constants.

The lumped-constant branch-line hybrid is adjusted by individually resonating the inductors with the shunt capacitances present at each port. Capacitors are well marked and can also be easily and accurately measured; therefore, only the inductors are trimmed. Values close to the calculated capacitance found in **table 1** are soldered in shunt with each port. Inductor L1 is then soldered from port 1 to port 2 to form a π network, as shown in **fig. 2A**. The coil is adjusted to resonance according to:

$$f = 1/\left(2\pi\sqrt{L1 \cdot C1/2}\right) \qquad (5)$$

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fig. 2. Experimental results using a lumped-constant hybrid are very close to the theoretical values (solid line). *2A* shows input return loss vs. frequency; *2B* shows isolation vs. frequency; *2C* shows coupling versus frequency.

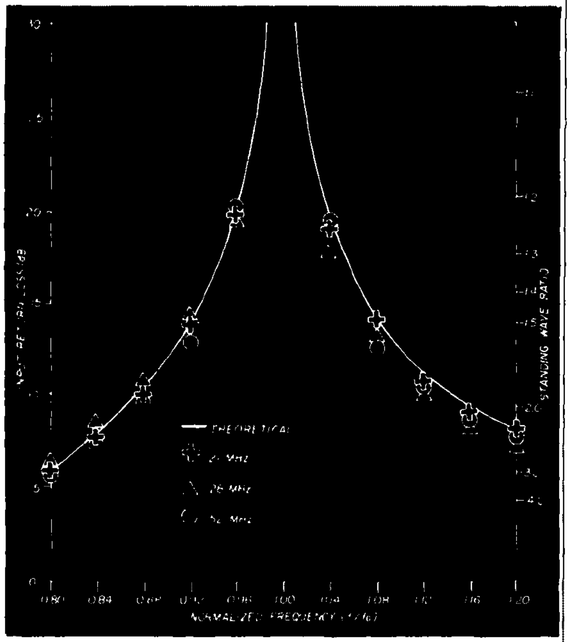

table 1. Lumped constant values for branch-line hybrids.

frequency (MHz)	L1 (nH)	L2 (nH)	C1 (pF)
3.75	1500.0	2122.0	2049.0
7.15	787.0	1113.0	1075.0
14.18	397.0	561.0	542.0
21.25	265.0	375.0	362.0
28.85	195.0	276.0	266.0
52.00	108.0	153.0	148.0
146.00	38.5	54.5	52.6
222.00	25.3	35.8	34.6

The *C1/2* term indicates that the *C1* capacitors are in series, with the center tap grounded and the inductor floating. For example, the series inductor in a 28-MHz hybrid should be adjusted to resonate at 31.2 MHz. Next, L1 is removed and L2 is placed between ports 2 and 3. A grid-dip meter is again used to resonate L2 with the series combination of the shunting capacitors. For the 28-MHz hybrid, the resonant frequency of the shunt arm is 26.2 MHz. After each coil is individually resonated, all are replaced, and the hybrid is ready for use. The experimental results shown in **fig. 2** were achieved by adjusting the hybrids in the above manner.

impedance transformation

Often it is required for a signal source to deliver two equal signals at loads other than 50 ohms. The input and output impedances of an RF transistor power amplifier, for example, are in the 1 to 10 ohm region. This impedance transformation may be incorporated directly into the branch-line hybrid. The series arm, as before, is made equal to the square root of the input impedance times one-half the output impedance. The

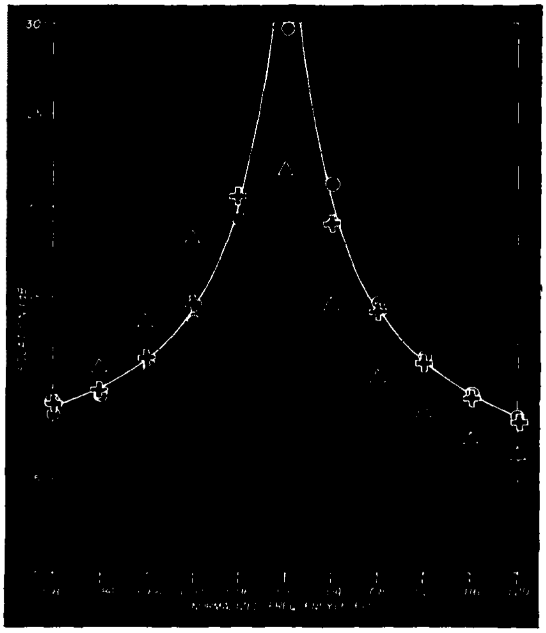

fig. 2B.

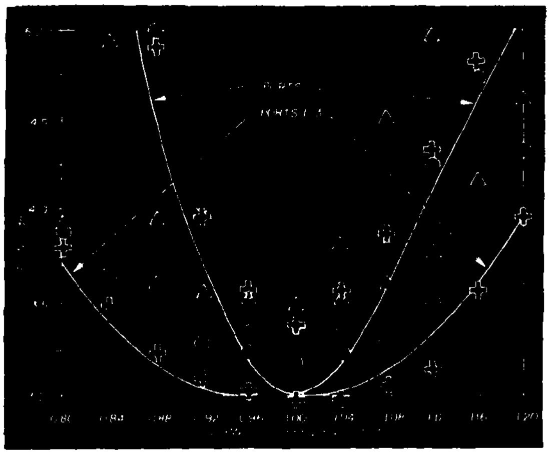

fig. 2C.

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one-half term occurs because the two output ports (2 and 3) are in parallel.

impedance ratio (ohms)	series arm (ohms)	shunt arm (ohms)
50/10 (5:1)	15.8	10.0
50/12.5 (4:1)	17.7	12.5
50/16.7 (3:1)	20.4	16.7
50/25 (2:1)	25.0	25.0
50/50 (1:1)	35.4	50.0

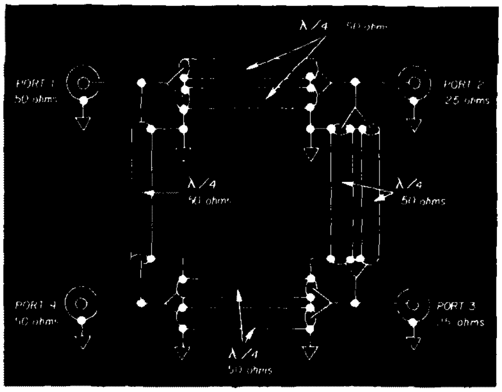
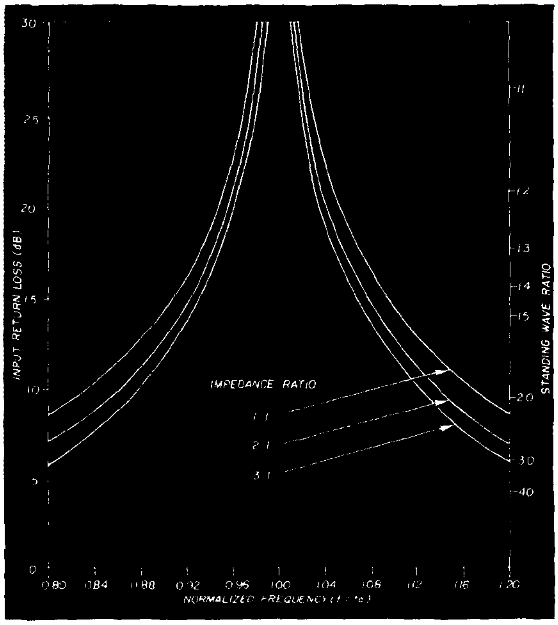


fig. 3. Common 50-ohm transmission line can be used to form a 2:1 impedance matching branch-line hybrid.

fig. 4. Theoretical response of transmission line branch-line hybrids used to transform 50 ohms to a lower value. *4A* shows input return loss versus frequency; *4B* shows isolation versus frequency; *4C* shows coupling versus frequency.



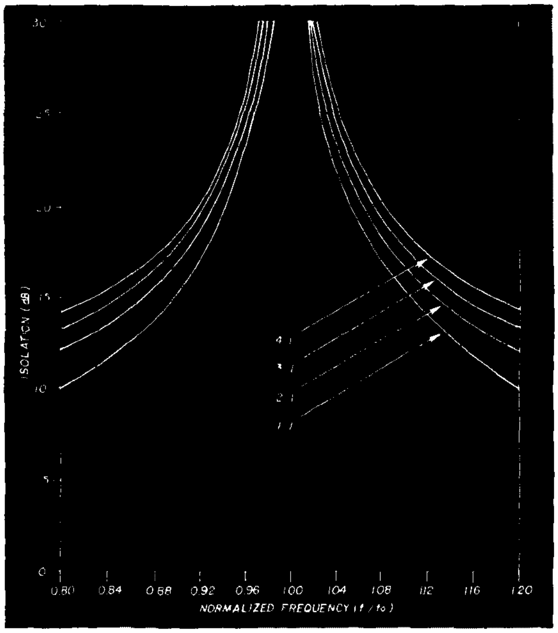


fig. 4B.

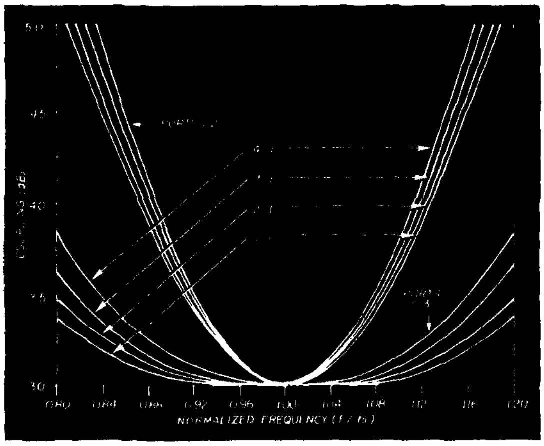
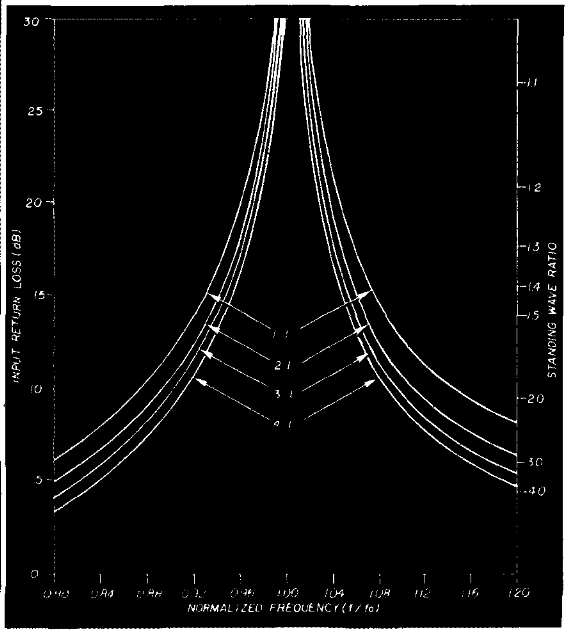


fig. 4C.

A convenient form of a 2:1 (50-ohm to 25-ohm) branch-line, impedance-matching hybrid is shown in **fig. 3**. The theoretical responses for several 50-ohm, input impedance matching branch-line hybrids are shown in **fig. 4** for various impedance transformation ratios. The response does not degrade significantly, even for an impedance matching ratio of **4**:1.

Lumped constants can also be used to form artificial transmission lines for an impedance transforming branch-line hybrid. The theoretical response, (**fig. 5A**),

fig. 5. Theoretical response of impedance-matching hybrids using lumped-constant components. *5A* shows input return loss versus frequency; *5B* shows isolation versus frequency; *5C* shows coupling versus frequency.



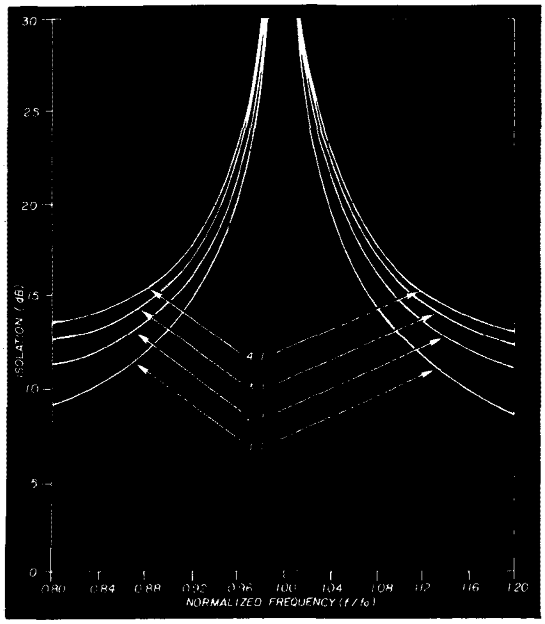


fig. 5B.

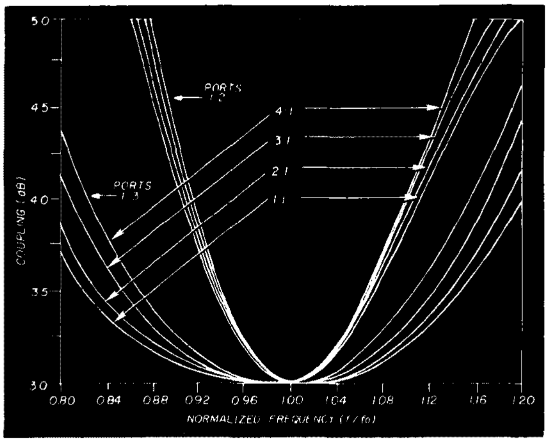


fig. 5C.

is only slightly degraded from that achieved using coaxial cable.

PIN-diode attenuator

The PIN-diode attenuator circuit,[3] **fig. 6**, is included to demonstrate that the input match remains constant despite large variations in the loads placed at ports 2 and 3. The resistance of a PIN diode decreases with forward bias. With zero bias current, the coupler behaves as a terminated hybrid with maximum attenuation from port 1 to port 4. As the diodes are forward biased, the resistance decreases. When each diode has a resistance of approximately 50 ohms, the attenuator has a loss of 10 dB, **fig. 7**; the diodes represent a mismatch. The reflected signals at ports 2 and 3 add in phase at the output, port 4, and out of phase to

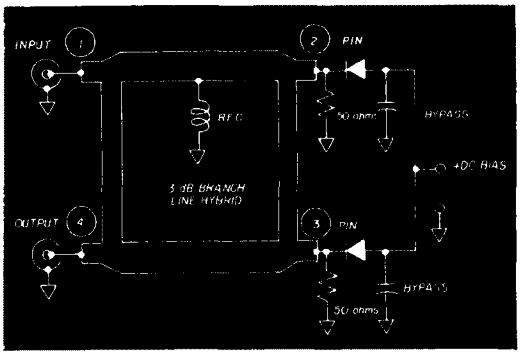


fig. 6. PIN diode attenuator with a constant input impedance of 50 ohms.

cancel at the input, port 1. As the bias is increased the loss of the attenuator decreases. When the PIN diodes' resistance is about 3 ohms, the loss decreases to 1 dB.

intermodulation performance

Even during a transmission the antenna operates as a receiving antenna for interference. Signals very close to the operating frequency, F_O, easily pass through

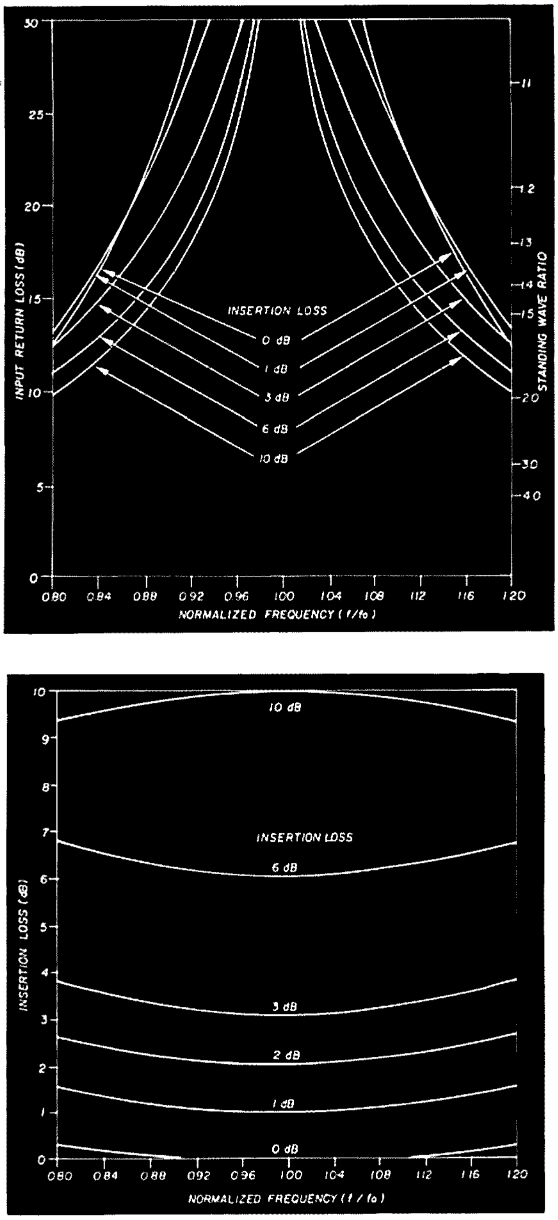
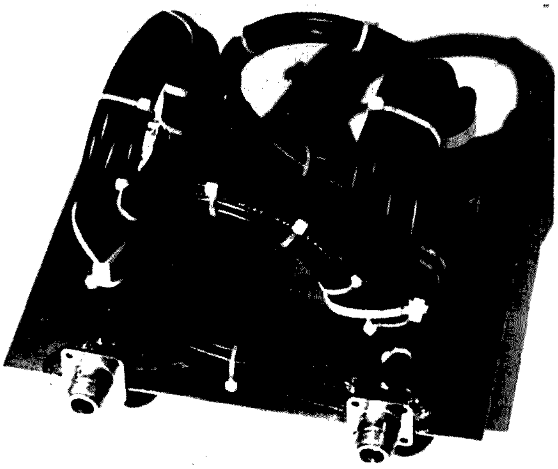
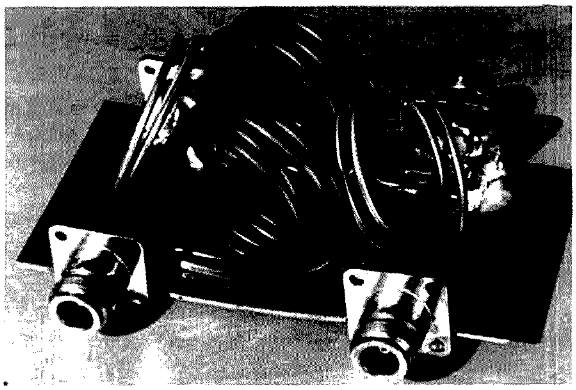


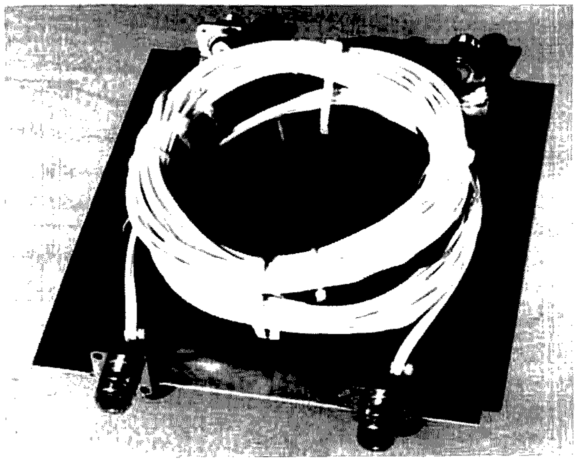
fig. 7. Input return loss remains greater than 15 dB (SWR $\leq$ 1.5) as the loss is varied between 0 and 10 dB.



At VHF frequencies, hybrids can be easily formed using several quarter-wave sections of 50 and 75-ohm coaxial cable.



For lower frequencies the semi-rigid cable must be coiled to contain the hybrid.



At low frequencies the loss in miniature coaxial cable increases the insertion loss.

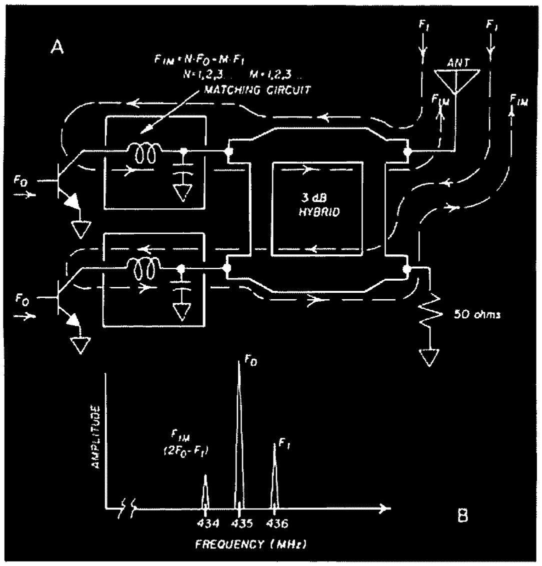


fig. 8. Intermodulation performance of a transmitter may be improved by adding a branch-line hybrid.

filters and arrive at the collectors of the final RF power transistors, **fig. 8(A)**. Because the final stage is operated Class C, mixing products are produced at these nonlinear transistor junctions. The most troublesome intermodulation product is produced by mixing the interfering signal, F_I, with the second harmonic of the operating frequency, **fig. 8(B)**. That is,

$$F_{IM} = 2F_O - F_I \qquad \textbf{(6)}$$

where F_{IM} is the intermodulation signal.

Because the collector ports of the branch-line hybrid[4] are offset by 90 degrees, the interfering signal must travel a quarter-wave longer from the antenna port to reach one collector compared with the other collector. The intermodulation product produced at the collector must also travel an additional quarter wave to reach the antenna port. Thus the intermodulation products from each collector will cancel at the antenna port because of the half-wave difference in round-trip paths.

references

1. James R. Fisk, W1HR, "Microstrip Transmission Line," *ham radio*, January, 1978, page 28.
2. Ernie Franke, WA2EWT, "The Hybrid Ring," *ham radio*, August, 1983, page 50.
3. Henry H. Cross, W1OOP, "Low-Noise Preamplifiers with Good Impedance Match," *ham radio*, November, 1982, page 36.
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ham radio

build a simple wire plow

Save time, effort with easy-to-build ground system installation aid

Any Amateur planning to erect a vertical antenna thinks long and hard about the labor involved in putting in the necessary number of radials. When I helped W7IR put in his radials, we used a small trencher, a tool used for installing the pipes for underground sprinkler systems. This worked fine, but it dug a trench much larger than was necessary for No. 14 wire, and replacing the soil after the wire was laid in was tedious work.

I had heard of the self-propelled wire plows used to install radials for broadcast stations, and I had seen one used in the installation of underground telephone lines. These tools simultaneously cut a slit in the ground and lay the wire. I wondered if it would be possible to make a smaller version of a wire plow that could be towed by a car or a garden tractor.

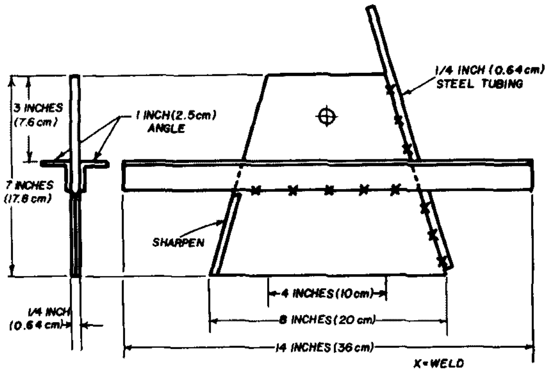

fig. 1. Construction details of wire plow blade.

My land was flat, reasonably unobstructed, and free of stones, at least to a depth of several inches. I convinced myself that the idea was practical. When I described my plans to friends at the local radio club, I got some skeptical comments, but my stubborn nature made me more determined than ever to go ahead. A little *pencil-scratching* produced a rough design, and the rest evolved as I went along.

Only the blade and some hardware had to be purchased. Everything else came from the scrap pile.

The blade, **figs. 1** and 2, was cut from 1/4-inch (6.4 mm) sheet steel by a local blacksmith for $10. I drilled the holes for fastening it to the frame and sharpened the leading edge on a bench grinder. For the wire guide, I bought a length of 1/4-inch (6.4 mm) steel brake line from an auto supply store. A friend with a welding outfit tack-welded it to the trailing edge of the blade (**fig. 2**).

radials

I planned to put in 36 radials, spaced at 10 degree intervals. These would vary in length from 80 feet (24.4 m) to 120 feet (36.6 m), keeping them inside the property line. I found that I could buy 2000-foot spools of No. 14 bare copper wire (**fig. 3**) from an electrical supply house.

The first thing to do was calculate the points at the edges of the property where the radials would terminate; then, with a 100-foot (30.5 m) tape measure and a friend's help, drive stakes at these points. The radials would be laid from the stakes toward the center, where a concrete pad had been poured for mounting the antenna.

plowing begins

I will admit that it was with some trepidation that the actual wire plowing was begun. I had invested a lot of time and labor in building the plow. Would it work?

At the first stake, W7IR and I dug a small hole, just large enough for the blade to enter the soil to its full depth. We drove a wooden stake into the hole and fastened the end of the wire to it (**fig. 4**).

I connected the plow to the trailer hitch on my car

By Harry R. Hyder, W7IV, 1638 Inverness Drive, Tempe, Arizona 85282

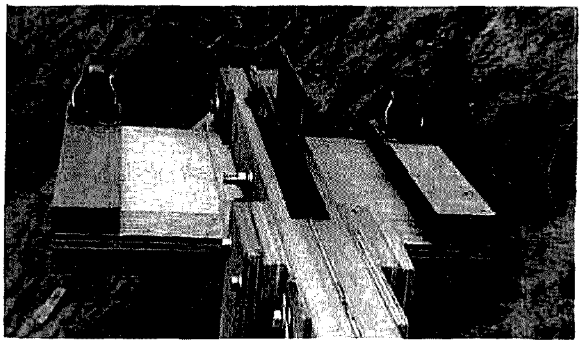

fig. 2. Bottom view shows blade, tack-welded to angle iron screwed into position. Wire guide tube is welded to trailing edge of blade (foreground).

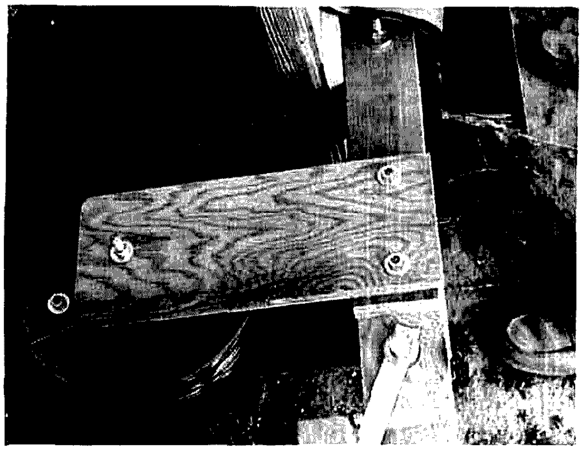

fig. 3. Wire is fed from mounted spool through guide tube into trench.

with a tow cable. W7IR steered the plow (**fig. 5**). I shifted the car into its lowest gear and inched forward. After a few feet I stopped, jumped out, and ran back to see how it as going.

It was working perfectly!

We laid half of the radials that morning, and completed the job the following Saturday. It took less time to install all the radials than it took to build the plow.

Only one minor problem emerged: during construction of my house, the builders had buried some chunks of concrete, and these had to be removed before the plow could proceed.

There was no reason to fill in the narrow slits left by the plow. Natural erosion took care of this, and after a few months they were barely visible.

securing the radials

In attaching the radials to the large sheet of copper on top of the concrete pad, I discovered that the copper sheet made a wonderful heatsink. My largest soldering iron — at the end of 200 feet of extension cable — could not solder the radials to the copper. A propane torch worked fine.

Another problem developed after the radials had been installed, but before the antenna was erected. The local telephone company, with their king-size wire plow, took a shortcut across the back of my property while laying a cable. While there was an eight-foot (2.5 m) easement running along the rear property line, specifically for utilities, they ignored this and cut several of my radials. (My radials did not extend into the easement.)

Afer working my way up through the telephone company bureaucracy, I was finally able to extract a promise that their cable would be relocated to its proper position; they did so a few months later. I chose to splice the severed radials myself.

Whether you can use a plow like mine depends on your circumstances. If your property is rocky and full of trees and brush, forget it. If it is clear with only light cover, the wire plow will work fine. It's a lot easier than using a pick and shovel!

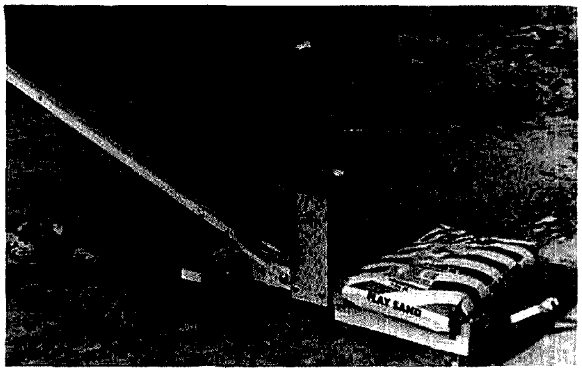

fig. 4. Wire plow in position over starting hole. Handle guides plow; 100-pound sandbag provides necessary weight.

fig. 5. W7IR, shown here, guided plow while author drove 4 × 4 in low gear; plow can also be pulled by garden tractor or conventional auto.

ham radio

VHF/UHF WORLD

Joe Reisert W1JR

In my February column,[1] I broached the subject of antenna performance but space did not permit a full treatment at that time; therefore, — and because this month's issue is dedicated specifically to antennas — I'll discuss several ways to determine antenna performance and offer some tips on how to obtain peak performance from the antennas you are now using.

performance parameters

When it comes to measuring antenna performance, most Amateurs measure only VSWR, and in the case of a rotary beam, front-to-back ratio. Consequently, commercial antenna manufacturers usually try to make sure that these parameters are good. All other specifications of a commercial antenna (or someone else's design you are duplicating) have to be accepted more or less on faith! This is unfortunate; you're at the mercy of the designer and may not be obtaining the performance you think you are.

Gain is surely one of the most important antenna parameters. However, other antenna parameters such as beamwidth, front-to-back ratio, side lobe level, VSWR, feed system, wind load and structural strength are likewise important and may be indirectly linked to gain. With this in mind, I will try to give you some guidelines to evaluate antenna performance as well as various measurements you can perform with a minimum of test equipment. Finally some rules of thumb and graphs will be presented that can help you approximate the gain and performance of your antenna or antenna system *even without test equipment*!

Gain. There is probably no other antenna parameter that is more widely discussed and confused by Amateurs than gain. *Gain is the property of an antenna which enables it to direct or radiate power in a desired direction as well as to receive signals from that same direction.* Note that when transmitting, an antenna doesn't really amplify or change the signal as in a power amplifier but does affect the direction of radiation similar to using a passive voice megaphone. Likewise, when receiving the antenna discriminates against noise and signals from undesired directions by not focusing on them. However, gain is a relative quantity that cannot be defined in terms of physical quantities such as watts or volts. Hence, gain must be referenced to something such as a dipole or an isotropic radiator (a theoretical antenna that radiates power equally well in every direction).

Herein lies the problem. *An isotropic radiator does not exist* (more on this later), but dipoles do! A lossless dipole has a theoretical gain of approximately 2.15 dB over an isotropic radiator, but a dipole is an extremely poor reference because it radiates power in many directions other than the desired one. This is why reflections from local objects, reflections from the ground, height above ground, and many other factors enter into the measurement of gain over a dipole.[2]

However, all is not lost! Over the years, VHF/UHF antenna manufacturers have developed acceptable gain standards that are accurate if properly used. The most common reference used by VHF/UHF Amateurs is the "EIA (Electronic Industries Association) Standard Antenna"[3] originally designed by Richard F.H. Yang.[4] It consists of two $\lambda/2$ dipoles which are spaced $\lambda/2$ apart and located $\lambda/4$ above a square groundplane with one-wavelength sides. It is easily duplicated and often used at antenna gain measurement parties (more on this subject later) on 432 and 1296 MHz and has a gain of 7.7 ± 0.15 dBd (dB over a lossless dipole) or 9.85 dBi (dB over an isotropic radiator). This antenna is often confused with the "NBS Standard Gain Antenna,"[5] which is considerably larger and has a gain of 9.31 ± 0.2 dBd. On the microwave frequencies accurate pyramidal standard gain horns[6] specified in dBi are usually used because they can be accurately designed and tailored to the desired reference gain. *It really doesn't matter whether we reference gain to a dipole or an isotropic radiator as long as we indicate what the reference is!*

Beamwidth. This is probably the most important parameter because it tells you how wide an area you are transmitting to and receiving from. It should be intuitive that in order to increase gain, the beamwidth of the antenna must become narrower. Wide beamwidth implies low gain and narrow beamwidth suggests high gain.

You cannot have high gain with wide beamwidth! It will be shown later in this column that the beamwidth parameter of an antenna can be used to mathematically determine gain.

Front-to-back ratio. This is a frequently quoted but somewhat elusive parameter. Yes, high front-to-back ratio does imply a high gain, but this is not necessarily the only important parameter for high gain. In fact, small (2 to 5 elements) Yagi antennas optimized for gain frequently have higher gain when the front-to-back ratio is only 10 to 15 dB.[7] Likewise, once the front-to-back ratio exceeds 20 dB, further increases will have little or no effect on gain or noise temperature since the rear signal is already so far down from the front one. It must also be remembered that the front-to-back ratio is measured over a small angle to the rear of the antenna and can be difficult to measure accurately due to its narrow angular width and reflections from other objects in your *local area.* High front-to-back ratio will, however, cut down QRM from strong local stations directly off the rear of an antenna.

Side Lobes. It wasn't that many years ago that side lobes on common Yagi type antennas were only 10 dB down from the main lobe. A typical "E" plane (azimuth) pattern on a widely used 144, 432, and 1296 MHz Yagi antenna is shown in **fig. 1**.[8,9,10] Unfortunately, the presence of side lobes on an antenna pattern is almost a fact of life. The higher the gain of an antenna, the more likely you are to have a greater number of lobes. *What is really important is how far down these lobes are with respect to the main beam, because they represent additional signal and noise pickup as well as lower gain.*

Fortunately the NBS Yagi designs[7] (an NBS 4.2λ Yagi's E plane pattern is shown in **fig. 2**) and other more modern techniques,[11] sometimes involving computers,[12] have improved design parameters, especially on Yagi antennas. Suffice it to say that it is desirable for serious work to have side lobes down at least 13 to 15 dB from

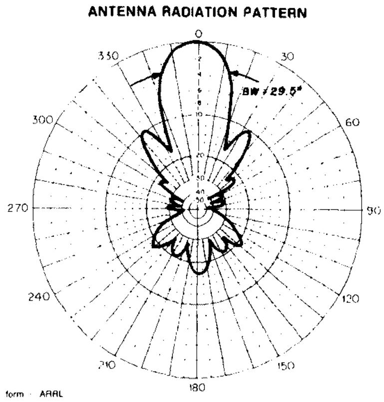

fig. 1. "E" plane plot of W2NLY/W6QKI 13-element 144 MHz Yagi. (See text and references 8, 9, and 10.) Note that beamwidth in "E" plane is approximately 29.5 degrees, and that the first side lobes are down only −10 dB. "H" plane beamwidth is the same (plotted on ARRL antenna radiation pattern paper.)

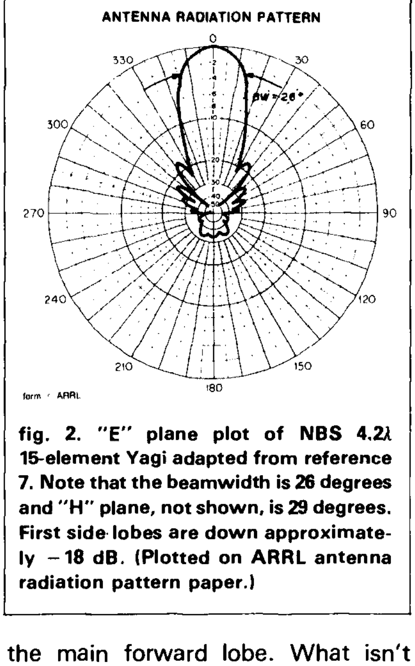

fig. 2. "E" plane plot of NBS 4.2λ 15-element Yagi adapted from reference 7. Note that the beamwidth is 26 degrees and "H" plane, not shown, is 29 degrees. First side lobes are down approximately −18 dB. (Plotted on ARRL antenna radiation pattern paper.)

the main forward lobe. What isn't always considered is that there may be numerous side lobes and therefore *it is desirable to have all side lobes — not just the first one — down as far as possible, especially for EME operation.*

VSWR. Antenna experts may tell you that you don't have to have a good VSWR to achieve full performance. However, VHF/UHFers have some slightly different problems than HFers do mainly because of mismatch losses. Let me be more specific. In the typical VHF/UHF station, 1-2 dB transmission line losses are quite common. If your antenna VSWR is 3:1, there will be an additional mismatch loss of 0.5 to 0.8 dB.[13] Therefore, your actual transmission line loss will be 1.5 to 2.8 respectively! Good VSWR is particularly desired when two or more antennas are stacked and fed with an in-phase power divider. Furthermore, a mismatch can frequently cause the noise figure of a low-noise preamplifier to increase since the preamplifier is no longer "seeing" the optimum source impedance it was designed for. Fortunately this parameter is seldom a problem nowadays, since low VSWR (1.2:1 typically) is quite common on most VHF/UHF antennas in use.

Feed System. Most VHF/UHFers agree that for best performance, an antenna should have a balanced feed system. Gamma and similar unbalanced matching systems can cause problems such as radiation from the feedline or assymmetry in the antenna pattern — two things we don't want! In the last few years, I have been doing some testing on various feed systems for Yagi antennas. Over all, I feel that the optimum feed system is a "T" match with a built in λ/2 coaxial type balun. Feed systems and baluns are covered in detail elsewhere[14] so I will not spend any more time on them in this column.

Wind Load. The modern VHF/UHFer is starting to challenge the HFers when it comes to antenna size. As a result, more attention has to be paid to wind survivability. Suffice it to say that *the optimum antenna should have the least number of elements necessary to attain the electrical specifications,* especially when stacking is employed. Don't forget that the wind load will increase rapidly when a large diameter stacking frame is used. In this respect, it may be prudent when selecting Yagi antennas to use the fewest number of long Yagis rather than a larger number of smaller models. Only by calculating the overall wind load on the array[15] can

you be sure that your tower and rotator will adequately support the antenna.

Structural Strength. It wasn't too long ago that VHF/UHF Yagi antennas were almost always constructed with heavy-wall tubing. Nowadays it is becoming more popular to use various lengths of thin-wall tubing. As a result, more attention must be paid to tubing diameter, wall thickness, tubing overlap, and the hardware used to connect sections of tubing. Often overlooked is the fact that a 25-30 percent smaller width square tubing may have the same strength as its round cousin. Square tubing is usually much easier for the homebrewer to drill. Trusses are strongly recommended on long antennas especially where wind and ice are prevelant. Never skimp *when choosing adequate size* bolts and nuts with lockwashers. Stainless steel hardware, despite its initial higher cost, is highly recommended because it will not rust out at the least desirable moment!

Preventive Maintenance. One area in which I think we are all lax is in preventive maintenance. Time should be scheduled at least once a year for checking all bolts and nuts for tightness as well as cleaning corroded joints. All connectors should be checked for moisture or corrosion as well as coax shield integrity. I've had good luck keeping water out of connectors by using "Coax Seal."® One problem I've noticed over the years is "center pin creep," particularly on "N" and "LC" connectors, especially when extreme temperature variations are encountered. After connectors are in place for a while, and definitely where some pulling or twisting is present such as on a cable going around a rotator, the center pins may push forward or pull back. At the least opportune time the pin may disengage or break the connector pin to which it is mated.

There are two reasons why I've covered so much background material in this month's column. The first is that I wanted you to be aware of all the areas that should be considered in the selection of a VHF/UHF antenna, whether it be homebrew or purchased. The second reason is that certain parameters can be used to determine other parameters. Furthermore, in some cases you may be able to substitute tests or just evaluate a data sheet parameter to determine other performance data on an antenna *but only if these parameters are well understood*! Read on.

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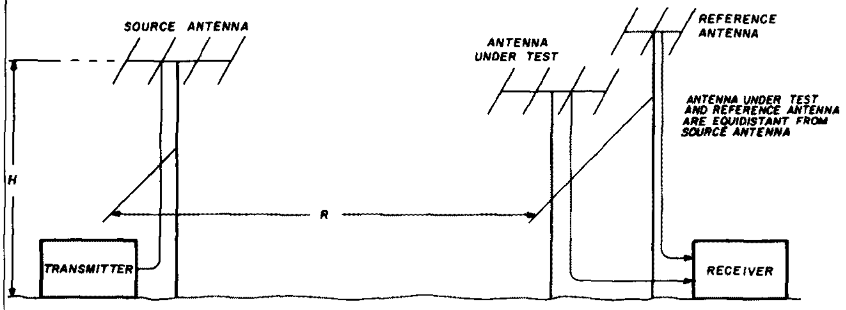

fig. 3. Typical VHF/UHF antenna test range. (See text for explanation of R and H test equipment.)

making measurements

Antenna Gain. This is probably the parameter you are most interested in testing. Gain can be measured on an antenna range. (A typical antenna range is shown in **fig. 3.**) A low-power transmitter is set up at one end and the antenna under test at the other end of the range. The transmitter is usually low power (1 watt maximum) with 1 kHz amplitude modulation. Typical circuitry has been published [16] for 144 MHz, and other frequencies through 432 MHz can be generated using the circuitry in my column in the March, 1984, issue of *ham radio* [17]. Note that the distance between the transmitter and receiver, "R," should be at least $2D^2/\lambda$ where D is the largest aperture dimension of the antenna and λ is the free space wavelength in the same units as D. I find that this yields only a 1.0 dB accuracy and hence I recommend $10D^2/\lambda$, which should yield 0.2 dB or better accuracy. "H" is the height of the source antenna and is described in detail in reference 2.

Basically what takes place is that the antenna under test is compared to the gain of a reference standard or antenna with a known gain, as discussed earlier in this article. Then the difference in gain between the two antennas is added to or subtracted from the reference as required to obtain the true measured gain. Many factors must be taken into consideration, but they are all well documented elsewhere.[2,18]

I want to point out that many VHF/UHFers have become quite competent at measuring gain and their results have been quite instrumental in making antenna designers "more honest" when it comes to gain specifications. We are very fortunate here in the USA in that antenna measuring parties are becoming quite common. They are usually held at various conferences such as the Central States VHF Conference, the West Coast VHF Conference, the Eastern VHF/UHF Conference, and more recently, at the Dayton Hamvention (did I miss anyone?), *weather permitting* of course. Therefore, if you want to measure the gain of your favorite antenna, attend one of these gatherings and see for yourself how the tests are conducted, as well as how your antenna stacks up against the competition!

So, you say, how do I measure antenna gain without assembling all this test gear? I'm glad you asked. There

is a way you can obtain reasonable performance evaluation using your own equipment. You'll note that earlier I mentioned the importance of knowing the half-power (−3 dB) beamwidth of your antenna because it is the most important parameter in terms of actual antenna gain. Measuring the beamwidth of an antenna is not a difficult task if you have a rotator with reasonable relative accuracy in the azimuthal plane.

It almost takes longer to explain the test procedure used to measure beamwidth than it takes to do the actual test. The idea is to first peak your antenna on a test signal such as a local Amateur *not on an obstructed path*, preferably near the frequency of interest (± QRM!). Note this reference level on your receiver "S" meter. Then rotate the antenna until the signal drops by 3 dB or about one half an "S" unit. This should be done as carefully as possible to maintain accuracy. Carefully note the direction. Now rotate the antenna toward the other direction, through the peak, and note the half-power point on the other side of boresight (boresight as in "straight ahead," and in this case, maximum response). Subtract the readings to obtain the true half-power beamwidth of the antenna under test. Then check how far down the first side lobes are and note this value for future reference.

The easiest way to insure "S" meter calibration accuracy of the half-power point is to place a 3 dB attenuator pad (as described below) in your IF line. (Note that attenuator accuracy is usually easier to obtain at lower frequencies). However, you can put the attenuator at the antenna input to your receiver if you have an accurate pad for the frequency of interest. For more accurate results, it is better to use three separate 3 dB pads, one ahead of, and one behind the reference pad as shown in **fig. 4**. This insures that the 3 dB reference is relatively unaffected by any impedance mismatches on the line.

An improvement in accuracy can usually be obtained by reversing the procedure as follows: first place the reference 3 dB attenuator in the line per **fig. 4** and peak your antenna on the signal of interest, noting the reading on your "S" meter. Next remove this attenuator. The signal will increase, hopefully by 3 dB, but the reading is not important. Next rotate the antenna to either side of boresight until the meter returns to the original value before the attenuator was removed. The rotator directions indicated are the −3 dB points.

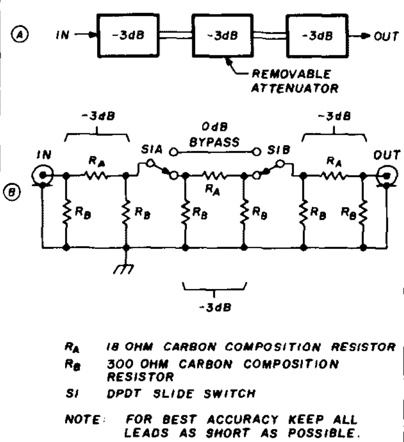

fig. 4. Recommended calibration attenuator: (A) with coaxial attenuator pads, and (B) with homebrew switchable attenuator.

If this test is too difficult or too time-consuming to perform, you may want to try a less accurate but faster method. Basically you modify the above procedure this way: first peak the antenna on the test signal and then rotate the antenna in either direction for the first null. Note the direction of the null and rotate back through boresight for the corresponding null on the other side of the pattern. Subtract the rotator readings. The antenna 3 dB beamwidth is approximately 47.5 percent of the difference between nulls.[19] For example, if the first nulls are 60 degrees apart, the approximate 3 dB beamwidth is 28.5 degrees.

Finally we're ready to apply your test results to see what the gain of your antenna really is! Many years ago John Kraus, W8JK, pointed out that the gain of an antenna could be roughly approximated if the "E" and "H" beamwidths of an antenna were known.[20] He also noted that the method was accurate only if the antenna beamwidth is narrow (less than 27 degrees), the side lobes are low (greater than 30 dB down), and the feed system is highly efficient. The equation he derived is:

$$G = \frac{41253}{\theta_E \times \theta_H} \quad (1)$$

where G is the directivity gain as a numeric (Gain in dB = 10 $\log_{10}G$) over an isotropic antenna and Θ_E and Θ_H are the antenna −3 dB beamwidth in degrees in the horizontal and vertical plane.

Over the years this equation has been widely accepted by the professional antenna community where −20 to −30 dB side lobes are quite common. This formula also seems to be usable over a much greater range of beamwidths than originally intended with surprising accuracy if the side lobes are low. I have noticed that if you use the following equations, you can also account for side lobes and thus improve accuracy even further.

$$G = \frac{38400}{\theta_E \times \theta_H} \quad \text{for } -25 \text{ dB side lobes} \quad (2)$$

$$G = \frac{35000}{\theta_E \times \theta_H} \quad \text{for } -20 \text{ dB side lobes} \quad (3)$$

$$G = \frac{32600}{\theta_E \times \theta_H} \quad \text{for } -15 \text{ dB side lobes} \quad (4)$$

$$G = \frac{30000}{\theta_E \times \theta_H} \quad \text{for } -10 \text{ dB side lobes} \quad (5)$$

To save you some time and math I have drawn a graph (**fig. 5**) which includes not only the answers to **eq. 1** directly in dBi, but also includes the answers to eq. **5** for −10 dB side lobes. You can easily interpolate for intermediate side lobe levels or just use the appropriate equation above.

Let's take a few examples to see how this system works. In **fig. 1**, we see that the beamwidth of this once-popular Yagi antenna is approximately 29.5 degrees and the side lobes are down approximately −10 dB. Now locate 29.5 degrees on the graph in **fig. 5**. If we were to ignore the side lobes (as the original authors did in reference 8), and use the upper line, the gain would be 16.8 dBi or 14.65 dBd. However, if you account for the side lobes per **fig. 5**, and therefore use the

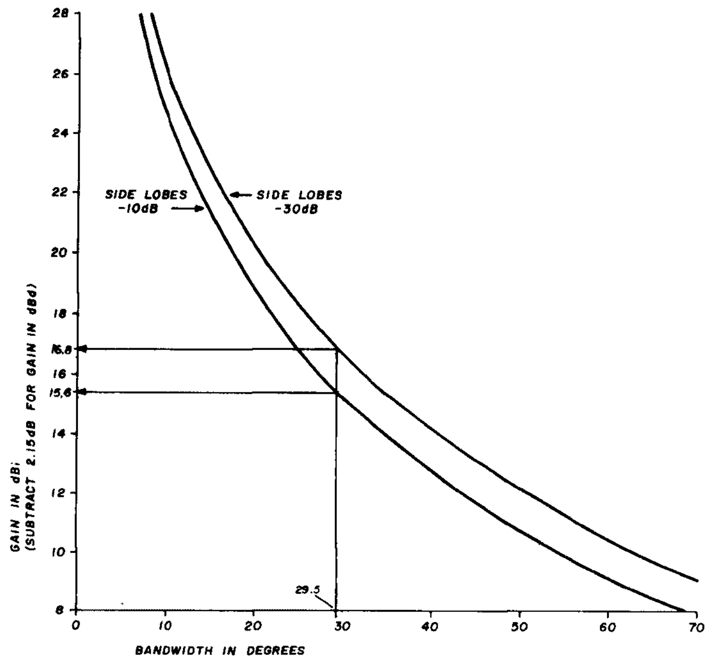

fig. 5. Beamwidth versus gain for antennas with square aperture and −30 dB side lobes or −10 dB side lobes. (See text for example shown.)

lower line on the graph, the gain is only 15.4 dBi or 13.25 dBd, a gain figure that was often measured on this Yagi on a good antenna range.

Next let's look at the NBS 4.2 λ Yagi in **fig. 2** which has a 26 degree beamwidth and −18 dB sidelobes. Using the upper line on the graph for no side lobes we see a gain of 18 dBi or 15.85 dBd but using **eq. 4** for 15 dB side lobes we obtain 16.83 dBd or 14.58 dBi. So you say NBS claims 16.35 dBi or 14.2 dBd? Well, I've left out one small factor. Some antennas do not have the same beamwidth in both planes. The error, however, is usually slight for antennas with rectangular apertures such as the Yagi. The "H" plane beamwidth for the NBS 4.2 λ Yagi is actually 29 degrees. If we go back and recalculate these numbers using **eq. 4** we obtain a gain of 16.35 dBi or 14.2 dBd. Not bad accuracy for just using simple equations!

As you've probably surmised, you needn't go through all this testing if you have accurate antenna data. In this respect, the commercial antenna manufacturers and Amateur designers usually include beamwidth data since it is easy to measure accurately. Now you can check the gain figures claimed and see how honest they are! (Always make sure that you use the correct data since some sources quote *half-beamwidth*. In this case, just double the number and proceed as shown above. And remember, is the gain quoted in the optimistic *dBi* figure or the lower *dBd*?

Other Testing. The front-to-back ratio can be tested using the methods above but may be influenced by local reflections and the accuracy of your rotator. If it is over 15 to 20 dB down from the main lobe, it is probably acceptable. VSWR, on the other hand, should always be tested before the antenna is raised to its final resting point as mentioned in the next section. Use a good VSWR indicator, not the "Monimatch" type. I recommend that you buy or borrow a Bird model 43 or equivalent with the appropriate power slug. If VSWR is not 1.2:1 or better, try to adjust the match until the optimum is attained. If it's above 1.5:1, you probably have trouble and better find the problem and fix it before placing the antenna in its final location!

Caveat Emptor. Sometimes commercial manufacturers make errors. This can sometimes be irritating even when the problem is simple and easily detected, such as in the case of a missing part. It's the undetected errors that cause real inconvenience — such as a wrong element length, for example, or a hole drilled in the wrong place in the boom. It is very easy to assemble directors on a Yagi improperly when the elements do not taper in a linear fashion. Nowadays most antenna manufacturers include mechanical drawings with element lengths and spacings clearly marked. *After final antenna assembly check carefully to see that all dimensions agree exactly!* If they don't, try to locate the source of the problem. If it is not obvious, contact the manufacturer and *get the problem resolved before you put the antenna on a tower.*

Let me share with you some problems I've encountered over the years so that you'll be alerted to things that can and do happen. One manufacturer copied another's design, forgot to correct the element lengths for a different element attachment method, and therefore had very high side lobes. Another drilled a boom improperly and thereby placed one director several inches off the proper location, causing high VSWR and poorer pattern. Another apparently had a mechanical resonance in the boom. The forward boom section would vibrate at certain wind speeds and eventually fall off because of metal fatigue. This was easily corrected by placing some weight inside the boom to dampen the effect.

One manufacturer uses end caps on certain elements. Neglecting to use these or having them fall off after mounting causes VSWR to increase. Some manufacturers use connectors that are screwed in place; during assembly, check these to make sure they are properly tightened. Another recent problem was a connector plate that was reversed during manufacturing and hence caused the antenna phase to be reversed by 180 degrees. This is not a problem with a single antenna, but a null on boresight can occur when the new antenna is stacked with a properly built model.

Most of these problems could have been caught prior to installation if the assembler paid attention to details, especially by cross-checking with the mechanical diagrams enclosed with the antenna. A VSWR check with the antenna mounted a wavelength or more above ground could also have pointed out other problems. Still other anomalies could have been noticed by testing the beamwidth, side lobes, and other parameters, using tests just described.

summary

In the past this information on determining antenna performance has been scattered throughout many different articles, with the gain methods mentioned, seldom used by Amateurs, buried in the math or appendicies at the end, with little explanation. I hope that this column has provided some practical guidance and that you will use it to compare one antenna's performance against another's. I'll refer to this information in future columns; good luck in selecting or evaluating your antenna!

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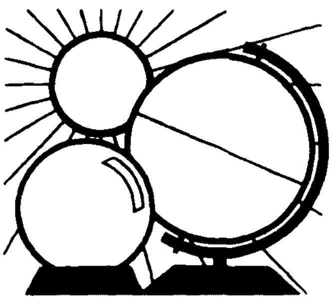
Important VHF/UHF Events in May, 1984:

May 2: ARRL 432 MHz Sprint Contest
May 4: 0730 UTC, predicted peak of Eta Aquarids Meteor shower
May 4,5,6: Tenth Annual Eastern VHF/UHF Conference, Sheraton Tara, Nashua, NH. (Contact K1LOG for further information.)
May 5,6: West Coast VHF Conference, Paso Robles, CA. (Contact K6HXW, Box 493, Arroyo Grande, CA 93420.)
May 10: ARRL 1296 MHz Sprint Contest
May 12,13: EME Perigee weekend
May 19: ARRL 6-meter Sprint Contest

ham radio



DX FORECASTER

Garth Stonehocker, K0RYW

sporadic-E propagation

In summer the overhead sun fills the lower ionosphere with ions which support short skip propagation, even multiple short skips. The geomagnetic field clusters these ions into cloud-like patches known as sporadic-E (E_S). To make best use of E_S DX openings, which are enhanced from late May until mid-September, a short review is in order.

E_S is a thin layer of intense ionization about 60 miles (100 km) above the earth. It gives rise to strong, mirror-like signal reflections over short-skip distances of 600 to 1200 miles (1000 to 2000 km). Signals remain strong for about a half-hour up to a couple of hours after the onset of the first strong signals, on the average; they're generally stronger than long-skip. Station location also determines how strongly the sunspot/geomagnetic disturbances affect sporadic-E propagation, with mid-latitudes the least affected and equatorial and polar paths the most. The highest *frequency* propagated by E_S occurs at local noon, since it follows the sun across the sky. However, the highest *probabilty of occurrence* is near sunrise and again around sunset. These two characteristics of E_S affect short-skip openings differently. Openings on the higher-frequency bands occur near local noontime; the lower bands tend to have openings near sunrise and sunset.

Because E_S is related to the summer sun, the best locations for these E_S openings are in the Northern Hemisphere from June through September and in the Southern Hemisphere during their summer, December through March. The best E_S is on either side of the geomagnetic equator; it's especially good where the geomagnetic equator is furthest from the geographic equator and during geomagnetic disturbances. These special areas are Southeast Asia in the Northern Hemisphere and South America in the Southern Hemisphere, with the former the better of the two.

To look for E_S openings on the higher-frequency bands, monitor beacons on 6 and 10 meters and CB channel 19. Also monitor unused TV channels 2 through 5 for 6- and 2-meter openings. The lower bands don't need beacon monitoring because E_S openings (sunrise and sunset) are available most nights.

last-minute forecast

A slight solar flux increase, with the possibility of a few flares about May 18th, should enhance DX on the higher-frequency bands (10-30 meters) the third week of the month. During the first and last weeks, look to the lower frequency bands (40-160 meters) for the best DX. Long periods in which the geomagnetic field will be disturbed are expected around the 1-9, 13, 23-26, and 31st.

An annual eclipse of the sun begins on May 30 at 1354 UTC near Hawaii, but stretching from the equator to Canada. It then crosses the Americas to Europe, stretching from northwest Africa to Norway, where it ends at 1935 UTC. Maximum duration is 62 seconds as it swings through the path. You might try some DX propagation experiments and compare your results to propagation or DX the day before and after.

The lunar perigee and full moon, of interest to moonbounce DXers, occurs on the 12th and 15th of this month. An Aquarid meteor shower of interest to meteor-scatter and meteor-burst DXers peaks between May 4th and 6th, with rates of 10 and 25 per hour for the northern and southern hemispheres, respectively.

band-by-band summary

Six meters will provide occasional openings to South Africa and South America around local noontime by short-skip E_S.

Ten and fifteen meters will have many short-skip E_S openings, and long skip during high solar flux to most areas of the world during daylight. Some trans-equatorial openings associated with disturbed ionospheric conditions may occur in the evening hours.

Twenty, thirty, and forty meters will have DX from most areas of the world during the daytime and into the evening hours almost every day, either long-skip to 2500 miles (4000 km) or short-skip E_S to 1250 miles (2000 km) per hop. The length of daylight is now approaching maximum, providing many hours of good DXing.

Thirty, forty, eighty, and one-sixty meters are all good for nighttime DX. On many nights 30 and 40 meters will be the only usable bands because of thunderstorm QRN, but signal strength via short-skip E_S may overcome the static when E_S is available. Although E_S is scarce in May, it should occur often the following two months.

MAY

WESTERN USA

GMT	PDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	5:00	20	15	15	10	20*	10	10	15
0100	6:00	20	15	15	10	20	10	10	15
0200	7:00	20	15	20	10	20	10	10	15
0300	8:00	20	20	20	10	20	10	10	15
0400	9:00	20	20	20	10	20	10	10	15
0500	10:00	20	20	20	10	20	10	10	15
0000	11:00	15	20	20	15	20	10	15	15
0700	12:00	15	*20*	*20*	15	30	15	15	20
0800	1:00	15	*20*	*20*	*15*	30	15	15	20
0900	2:00	15	*20*	*20*	*15*	30	15	20	20
1000	3:00	20*	*20*	*20*	20	*30*	20	20	20
1100	4:00	*20*	20	*20**	20	*30*	*20*	20	*20*
1200	5:00	*20*	20	15	20	*30*	*20*	*20*	*20*
1300	6:00	*20*	20	15	20	*30*	*20*	*20*	*30*
1400	7:00	20	20	15	20	*30*	20	*20*	*30*
1500	8:00	20	20	15*	20	20	20	20	*30*
1800	9:00	20	20	10	20	20	20	20	20
1700	10:00	20	20	10	20	20	15	20	20
1800	11:00	20	20	10	15	20	15	15	20
1900	12:00	20	20	10	15	20	15	15	20
2000	1:00	20	20	10	15	20	15	15	20
2100	2:00	20	20	10	15	20	10	15	20
2200	3:00	20	15	15	15	20*	10	10	20
2300	4:00	20	15	15	10	15	10	10	20*

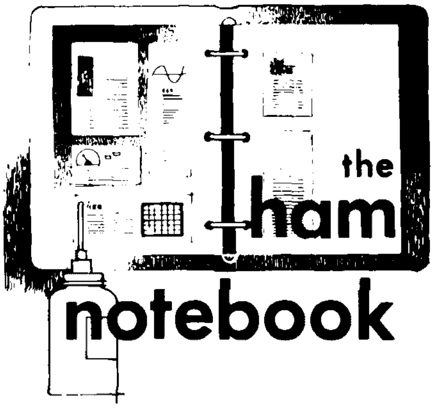
MID USA

GMT	MDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN	CDT
0000	6:00	20	15	15	15	20	10	10	15	7:00
0100	7:00	20	15	15	20*	20	10	10	15	8:00
0200	8:00	20	20*	20	20	20	10	10	15	9:00
0300	9:00	20	20	20	20	20	10	10	15	10:00
0400	10:00	20	20	20	20	20	10	10	15	11:00
0500	11:00	20	20	20	20	30	10	10	15	12:00
0000	12:00	*20*	20	20	20	30	10	15	15	1:00
0700	1:00	*20*	20	20	20	30	15	15	20	2:00
0800	2:00	*20*	*30*	*20*	*20*	30	15	15	20	3:00
0900	3:00	*20*	*30*	*20*	*20*	*30*	15	*20*	20	4:00
1000	4:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	*20*	5:00
1100	5:00	*20*	*30**	*15*	*15*	*30*	*20*	*20*	*20*	6:00
1200	6:00	*20*	20	15	15	*30*	*20*	*20*	*20*	7:00
1300	7:00	*20*	20	15	15	*30*	*20*	*20*	*30*	8:00
1400	8:00	20	20	15	15	*30*	*20*	*20*	*30*	9:00
1500	9:00	20	20	10	10	20	20	20	*30*	10:00
1800	10:00	20	20	10	10	20	20	20	20	11:00
1700	11:00	20*	20	10	10	20	15	20	20	12:00
1800	12:00	15	20	10	10	20	15	15	20	1:00
1900	1:00	15	20	10	10	20	15	15	20	2:00
2000	2:00	15	20	10	10	20	15	15	20	3:00
2100	3:00	15	20	10	10	20	15	15	20	4:00
2200	4:00	20*	20*	15*	15	20	10	15*	20	5:00
2300	5:00	20	15	15	15	20	10	10	20*	6:00

EASTERN USA

GMT	EDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE CARIBBEAN S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	8:00	20	20	15	15	20	10	10	20*
0100	9:00	20	20	15	15	20	10	10	15
0200	10:00	20	20	20	20	20	10	10	15
0300	11:00	20	20	20	20	20	10	10	15
0400	12:00	20	20	20	20	20	10	10	15
0500	1:00	20	20	20	20	30	10	10	15
0000	2:00	*20*	20	20	20	30	10	15	15
0700	3:00	*20*	*30*	20	20	30	15	15	20
0800	4:00	*20*	*30*	*20*	*20*	30	15	15	20
0900	5:00	*20*	*30*	*20*	*20*	*30*	15	*20*	*20*
1000	6:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	*20*
1100	7:00	*20*	*30*	*15*	*15*	*30*	*20*	*20*	*20*
1200	8:00	*20*	20	15	*15*	*30*	*20*	*20*	*20*
1300	9:00	20	20	15	15	*30*	*20*	*20*	*20*
1400	10:00	20	20	15*	15	*30*	*20*	*20*	*20*
1500	11:00	20	20	10	10	20	*20*	*20*	*20*
1800	12:00	20	20	10	10	20	20	20	*20*
1700	1:00	15	20	10	10	20	15	20	20
1800	2:00	15	20	10	10	20	15	15	20
1900	3:00	15	20	10	10	20	15	15	20
2000	4:00	15	20	10	10	20	15	15	20
2100	5:00	15	20	10	10	20	15	15	20
2200	6:00	20	20	15	15	20	10	15*	20
2300	7:00	20	20	15	15	20	10	10	20*

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during 'normal' hours.
*Look at next higher band for possible openings.



TS-430S IF filter mod

The Kenwood TS-430S allows for great flexibility in selecting different IF filters. It comes with a 2.4 kHz (6 dB) wide filter for SSB and CW and has a slot for both narrow SSB (1.8 kHz) and narrow CW (either 270 or 500 Hz) filters. In addition, a 6 kHz wide filter is available for AM.

The filters are automatically selected according to the mode selection (CW, USB, LSB, etc.) and according to the narrow/wide switch. One limitation of this is that the narrow CW filter cannot be used in the LSB mode. Most RTTY operation with this rig will be done in the LSB mode, and having a narrower filter would reduce the effects of QRN and QRM.

The original circuit (**fig. 1A**) uses a diode switching approach to select the appropriate filter. Connector No. 27 (may be labeled No. 29 in the instruction manual) comes from the narrow/wide switch on the front panel. Either of the control lines SSW or CWW go high in the wide position, depending on the mode (SSB or CW respectively). Similarly, SSN or CWN go high in the narrow position, according to the mode selected. These control lines connect through resistors (and sometimes diodes) to the appropriate IF filter. A modified circuit (**fig. 1B**) is shown which allows for the use of the narrow CW filter in the SSB mode. In the LSB mode, in particular, this filter ends up being centered around an audio frequency of about 2 kHz, which is ideal for RTTY interfaces using audio tones in that area. The IF shift control can be used to adjust this frequency, if necessary.

In normal transceiver operation, switching from CW or SSB often results in the narrow position selected in the SSB mode. If no narrow SSB filter is installed, the IF section is simply left open and no signals can be heard. The suggested but untested circuit shown in **fig. 1C** will enable the wide filter in SSB mode regardless of the narrow/wide switch position.

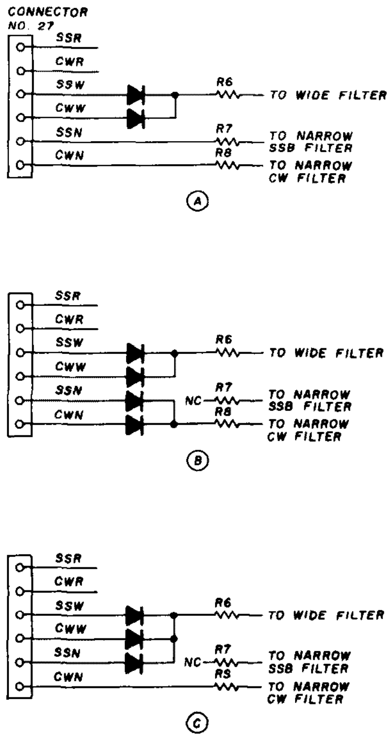


fig. 1. (A)Original circuit selects the appropriate narrow filter depending on whether the transceiver is in SSB or CW mode. (B)Modified circuit allows for narrow CW filter to be used in both CW and SSB modes. (C)Suggested circuit for enabling the wide filter in SSB/narrow mode.

To locate the appropriate area inside the transceiver, follow the instructions for installing the optional filters. R7 and R8 are located towards the rear of the IF filters. The ends of R7 and R8 nearest the rear of the transceiver go to connector No. 27. (Be careful in unsoldering the resistors because the printed circuit board traces are delicate.)

Robert A. Witte, KB0CY

short circuits

PL tone generator

In the April, 1984 article, "A Programmable PL Tone Generator," the component labeled "C7" in the upper left-hand corner (immediately below "U7") of the parts placement diagram (**fig. 5**, page 56) should be labeled "C17."

RF synthesizers

In the three-part series "RF Synthesizers for HF Communications," (August, September, October, 1983), three corrections are required. First, the labeling of the vertical axis in the phase noise plots in **figs. 15** and **20** of **Part 2** (September) and **fig. 13** of **Part 3** (October) should read "P_n dBc (IN 1 HZ BANDWIDTH)."

A second *error* appears in the last paragraph of the September article (page 50) and under "Using Bode Plot" on page 26 of the October issue. The open-loop unity gain frequency ($f_{\beta O}$) is not exactly equal to the closed loop 3 dB frequency (f_β). While it is a reasonable approximation, the two are more accurately related by:

$$f_{\beta O}/f_b = \frac{\sqrt{2\xi^2 + \sqrt{(2\xi^2)^2 + 1}}}{\sqrt{2\xi^2 + 1 + \sqrt{(2\xi^2 + 1)^2 + 1}}}$$

This difference is caused by the fact that at open-loop unity gain, the phase lag is not exactly 90 degrees. The closer this lag approaches 180 degrees (i.e., damping *decreasing*), the larger the difference between the open-loop unity gain frequency and closed-loop 3 dB frequency. (This ratio is approximately 0.755 at $\xi = 0.707$, 0.829 at $\xi = 1$, and 0.943 at $\xi = 2$.)

Finally, in **fig. 9** of **Part 3** (October) a 0.01 μF capacitor should be connected between pins 2 and 6 of the NE5534 op-amp (as described in the text).

Use GDO, noise or RF bridge to determine resonance

matching dipole antennas

This article describes how to use a grid dip oscillator (GDO), RF bridge, or noise bridge to measure an antenna's resonant frequency and its resistance at resonance. While the article specifically addresses dipole antennas, the procedure described is applicable to any antenna, and is meant for those Amateurs who wish to feed antennas directly without the need for a transmatch or other matching device.

cutting to formula

The starting point for finding a dipole's resonant length is:

$$L = \frac{468}{f} \tag{1}$$

where L = length in feet
f = frequency in MHz

This length allows for "end effects." Specific height above ground, length-to-diameter ratio and proximity to other objects (particularly conductors) also influence the resonant length.

As an example, a recently erected 10- and 15-meter dipole cut to formula had to be adjusted by removing six inches from one dipole and adding even more to the other. These are appreciable changes in 33- and 22-foot dipole lengths — about 2 percent. These length changes also represent about a 2 percent frequency variation: 280 kHz on 20 meters and 420 kHz on 15 meters. These changes are about as large as the bands are wide! Even though the resonant frequency of the antenna is "off," this information is still useful in determining correct length by working with "percentages." For example, if you find that the resonant frequency of the antenna *at its operating height* is 2 percent low, correct by reducing the overall length of the antenna by approximately 2 percent. Because this change is still relatively small, the 2 percent reduction can be made to one side only.

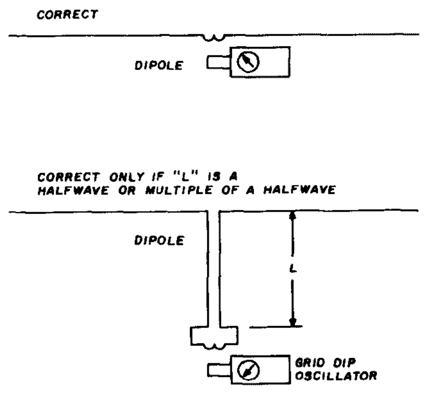


fig. 1. "Grid dipping" for the resonant frequency. Eliminating the feedline removes false resonances that can be more pronounced than the real one.

eliminating feedline ambiguities

If possible, it's best to make measurements at the antenna terminals directly (**fig. 1**). If you try to measure the resonant frequency of the antenna from the transmitter end of the cable, you'll see a combination of effects related to the antenna, feedline, and matching devices between the antenna and yourself.

However, if your feedline is a multiple of an electrical half-wave at your antenna's resonant (and desired operating) frequency, the feedline will reflect the antenna's characteristics at its input end (at the transmitter) (**fig. 1**). This way the antenna can be checked at its final height. To do this, determine the approximate feedline length using eq. **2**:

$$L_{\lambda/2} = \frac{492 \bullet V_F}{f} \tag{2}$$

where $L_{\lambda/2}$ is electrical half wavelength in feet
V_F is velocity factor (between 0 and 1)
f is frequency in MHz

Because the velocity factor of cables can vary from manufacturer to manufacturer, and even within the

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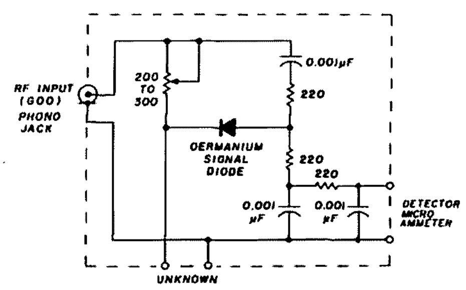


quantity	*description*
1	*enclosure RS-270-235*
4	*binding posts RS-274-661*
1	*phone jack RS-274-1575*
1	*potentiometer, 200 to 300 ohms carbon composition (typically, Allen Bradley type AB; Claristat, type 53; or Ohmite, type J)*
3	*resistors, 1/4 watt, 220 ohm 5 percent RS-271-1313*
3	*capacitors, disc 0.001 µF RS-272-126*
1	*germanium signal diode (IN34A) RS-276-1123*

Note: Radio Shack™ part numbers are shown

fig. 2. **RF bridge circuit.**

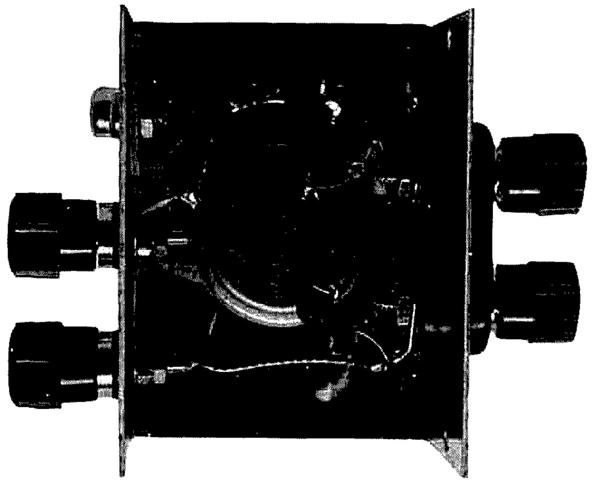


fig. 3. Interior view of the RF bridge shows the parts arrangement. The layout is not critical. Be sure to use non-inductive, carbon composition resistors. Terminals may be substituted to suit your test equipment.

same manufacturer's run of cable, more accurate approaches — such as a GDO, noise or RF bridge — should be used to determine the exact half wavelength of cable. The input end of a cable is resonant when it is shorted at a half or multiples of a half wavelength or when it is open-circuited at a quarter wavelength or at multiples of an *odd* quarter wavelength. These facts make a simple measurement possible.

To find an electrical half wavelength using a GDO, cut the cable a little longer and short its far end. Connect a small coupling coil (a single shorted turn might be sufficient) and bring the GDO up close. Use the smallest coupling coil possible to minimize error. Trim the far end as needed. (A more accurate frequency indication can be achieved by listening to the GDO on an accurately calibrated receiver.) If you want a quarter wave cable length rather than a half wave, repeat the procedure with half the length of cable, but do not short the far end.

When using a noise or RF bridge, the technique for determining a half wavelength of cable requires setting the bridge for zero resistance (and reactance if your bridge is equipped to measure reactance), cutting and shorting the cable as above and connecting the bridge to the input end of the cable. Resonance is indicated by a reduction in received noise in the case of the noise bridge or a minimum reading on an RF bridge's null detector. Trim the cable until the resistance at the input end goes to zero (or as close to zero as is practical to obtain). A quarter wavelength is determined in the same manner, but the far end of the cable is open-circuited.

The RF bridge and GDO combination can be used for outdoor antenna work because these devices are usually quite portable. Noise bridges themselves are easily portable, but the combination of bridge and receiver presents a real challenge to mobilty. On the other hand, noise bridges are simpler to set up and handle than the RF bridge and GDO combination. If you plan to do more complex antenna and/or RF circuit work, noise bridges will provide accurate reactance and resistance measurements over a very wide range of frequencies.

building an RF bridge

The bridge shown in **fig. 2** uses an outboarded 100 microampere meter for sharp and deep nulls. Parts placement is not critical, but a metal enclosure should be used to minimize "hand-capacity" variations. All parts are available from Radio Shack except the balance potentiometer, which should be a linear carbon composition type: typically, Allen Bradley, Type AB; Claristat, Type 53; or Ohmite, Type J.* Its value should be about 200 ohms. Most of the measurements will be in the order of 50 ohms. A low value potentiometer will assure that the scale is expanded enough to read easily (see **fig. 3**).

The bridge is calibrated using an ohmmeter. (1 percent resistors would provide greater precision — **Ed**). Mount a paper scale (see **fig. 4** and **5**) behind the pointer, marking each 10-ohm point between 10 and 100 ohms and each 50-ohm point above 100 ohms. Calibration may be done with the circuit wired up since

*Most electronic supply houses carry these types of potentiometers but have minimum order amounts of at least $20.00. Group ordering is a possibility; the potentiometers are in the $4 to $5 price range. Most large cities have one or two suppliers that will sell to Amateurs over the counter. One such is Linear Electronics in Waltham, Massachusetts.

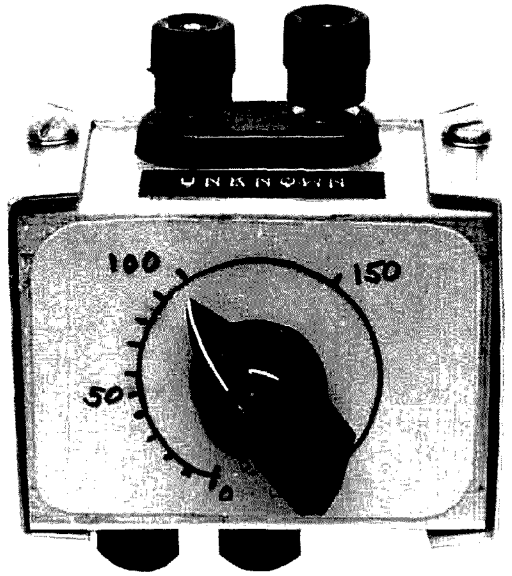


fig. 4. Panel view of the RF bridge shows the calibration of the variable (balance) resistor. This is done using an ohmmeter – no trick to it all! Make sure the variable resistor value increases in a clockwise direction.

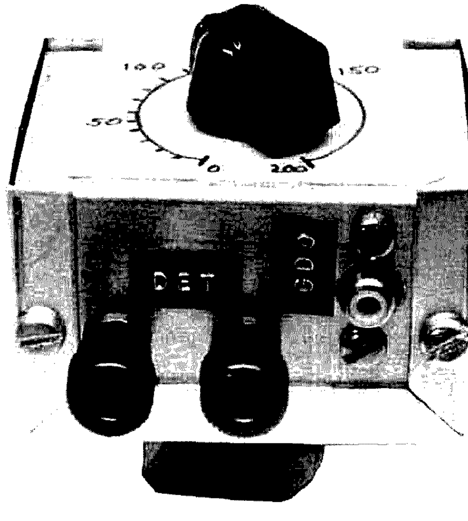


fig. 5. Top view of the RF bridge shows the jack for connecting the grip dip oscillator and the detector terminals.

the potentiometer is isolated by capacitors. Make sure the correct potentiometer ends are connected such that the scale increases in a clockwise direction.

The GDO is coupled to the bridge using a small coil and a short length of twisted hook-up wire. A two-turn loop slightly larger than the coil in the GDO is usually adequate. Tight coupling initially may make the dip easier to find, but afterward use as little coupling as practical to assure minimum interaction between the GDO and the circuit being measured.

using the bridge

Connect the bridge, null indicator (microammeter) and GDO as shown in **fig. 6**. The dipole should be connected as directly to the bridge as possible. Turn the GDO on and adjust the coupling for about a half-scale reading on the null detector. Set the resistance knob to about 50 ohms. Vary the GDO frequency until the detector microammeter, *not* the meter on the GDO, goes through a minimum (dip). Set the frequency for the lowest dip and then adjust the resistance knob to further increase the dip. Loosen the GDO coupling until the dip is barely perceptible and again set the frequency and resistance for the best possible dip. Now read the GDO frequency and bridge resistance. These are the antenna's resonant frequency and its resistance at that frequency, respectively. If corrections in length (shorter to increase and longer to decrease the frequency) are necessary, proceed as previously discussed and then recheck the frequency as just described. **Fig. 7** illustrates the use of a noise bridge and a receiver to determine antenna resonant conditions.

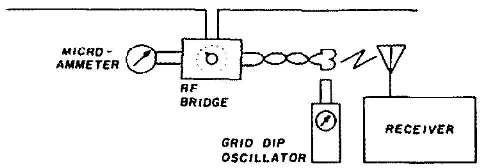


fig. 6. Measurement set-up for a simple RF bridge. The receiver is not necessary but may be used to increase the frequency measurement accuracy.

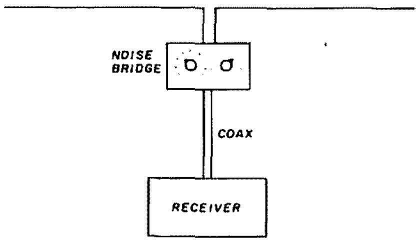


fig. 7. Measurement set-up for a noise bridge. The receiver senses less noise at the antenna resonance because it is being absorbed (transmitted) by the antenna when the bridge is properly balanced.

conclusion

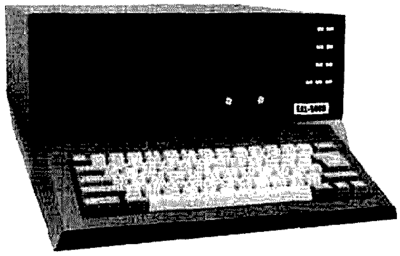
The material in this article should be particularly useful to those wishing to take their first steps toward antenna experimenation. More importantly, it may stimulate deeper interest in antennas and other tuned RF devices. The kind assistance of Domenic Mallozzi, N1DM, Robert Doherty, K1VV, and Clyde Shappee, KA1CRV, are gratefully acknowledged.

ham radio

NEW products

digital communications terminal

The EXL-5000 digital communications terminal from Amateur-Wholesale Electronics features a built-in high-resolution, long-persistence green monitor for sharp, clear images with no jiggle or jitter even under fluorescent lighting. Also featured are an external plug-in keyboard and a versatile built-in power supply for 117/234 AC or 13.8 VDC. Operation is greatly simplified by front-panel pushbutton controls.



The EXL-5000 includes capability for operating in AMTOR modes A, B, and L. A high-speed demodulator allows Baudot and ASCII operation from 12 to 300 bauds with AFSK and 12 to 600 bauds at TTL level, in increments of 0.1 baud. The demodulator will work with high or low tones. For transmitting, AFSK or FSK keying can be used. Morse sending and receiving speed is adjustable from 5 to 100 WPM in 1-WPM steps, with receive autotracking and variable transmitting weight.

A 1280-character display memory is split into two pages of 40 characters by 16 lines. Seven independently programmable 72-character channels and eight independently programmable 24-character channels allow storage of messages for permanent use. The memories can be preloaded and easily reprogrammed. Errors are easily corrected. The programmable memories are backed up by an internal battery so that they are never lost.

Other features of the EXL-5000 include full cursor control, function display, split-screen operation, automatic send/receive switching, automatic carriage return and line feed, automatic letters-code insertion, word-mode and line-mode operation, word wrap-around, simultaneous send-receive capability, selective calling, automatic timer-controlled transmission, RY and "quick-brown-fox" test signals, automatic ID, random-character generator for code practice, a printer interface, provision for an external monitor, a built-in audio monitor, a bar-graph LED tuning meter, noise-reduction receiving circuit, and time clock.

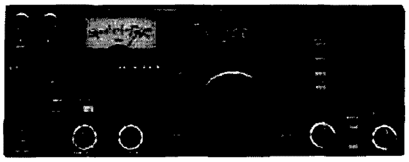
Further information about the EXL-5000 may be obtained from Amateur-Wholesale Electronics, Inc., 8817 S.W. 129th Terrace, Miami, Florida 33176.

Circle #301 on Reader Service Card.

IC-R71A

ICOM has introduced the new IC-R71A 100 kHz to 30 MHz professional-grade general coverage receiver, offering the same performance of the IC-R70 as well as several new features at the price of $799. This easy-to-use, versatile receiver features keyboard frequency entry, 32 programmable memories, SSB/AM/RTTY/CW/FM (optional), scanning, selectable AGC and noise blanker, passband tuning, and three tuning rates: 10 Hz/50 Hz/1 kHz. Two optional CW filters FL32 (500 Hz) and FL63 (250 Hz), are available, as well as an optional FL44A high-grade crystal filter (455 kHz).

The IC-R71A makes it possible for anyone, even without previous shortwave receiver experience, to listen to worldwide communications. Utilizing ICOM's DFM (Direct Feed Mixer), a 100 dB dynamic range, deep IF notch filter and adjustable AGC and noise blanker, the IC-R71A provides clear reception even in the presence of strong interference or high noise levels. A quartz-locked synthesized tuning system provides stable operation.



The pushbutton keyboard provides instant selection of frequencies which is accomplished by pushing the digit keys in sequence of frequency. Memory channels can be called up by pressing the VFO/M switch, then keying in the memory channel digit/s.

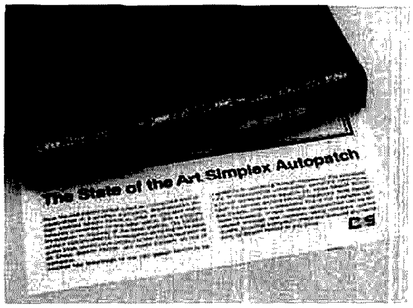
Options include FM, synthesized voice frequency readout, a wireless remote controller, a DC adapter for 12V operation, a mounting bracket, two CW filters and a high-grade crystal filter-455 kHz.

For further details, contact ICOM, 2112 116th Avenue, N.E., Bellevue, Washington 98004.

Circle #302 on Reader Service Card.

improved autopatch

A new simplex autopatch from CES will work on any Amateur or commercial simplex radio. CES engineers have redesigned the VOX enhancement circuitry and mobile presence detectors in the Model 510SA Smart Patch. The improvements allow the advanced microcomputer in the Smart Patch to keep the user from missing words or information. After trial testing at over 400 customer locations, it was concluded that the Smart Patch is now easier to install and is the only simplex patch that gives the mobile unit complete and immediate full break-in capability without loss of information. The immediate control feature allows operation in the Amateur service because Smart Patch cannot transmit on top of another mobile. Transmission can be terminated by simply keying the transmitter. Installation consists of connecting RX audio, TX audio, PTT and power.



For further information about the Smart Patch, contact Communications Electronics Specialties, Inc., P.O. Box 2930, Winter Park, Florida 32790.

Circle #303 on Reader Service Card.

feather-weight headset

Telex has introduced an ultra-light headset for hand-held land-mobile transceivers. The ProCom 352-IC weighs 2.6 ounces when worn *with* the headband, but only one ounce without, and can be clipped directly onto eye or sunglass frames for convenience. Priced at $129.95, the headset plugs directly into ICOM or Ten-Tec hand-held transceivers.

A soft ear tip channels incoming messages directly into the operator's ear so that communications are essentially private. The noise cancelling electret microphone is designed for very close talking and transmits the operator's voice clearly even in high-noise environments. The electret microphone is also immune to electromagnetic or RFI so it can be operated effectively near power lines, large transformers, generators, broadcast towers, and other equipment that often interferes with radio communications.

For more information, contact Telex Communications, Inc., 9600 Aldrich Avenue, South, Minneapolis, Minnesota 55420.

flea market

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Coming Events

ACTIVITIES

"Places to go..."

COLORADO: Rocky Mountain VHF Society's annual Spring Hamfest, Sunday, May 20, 9 AM to 3 PM, rain or shine, Boulder National Guard Armory, 4750 North Broadway, Boulder. Admission $3 per family, no seller's charge. Sellers please bring own tables. Tech demonstrations and seminars on packet radio, fast-scan ham TV, microwaves, satellite communications, etc. Food and drink available. Talk in on 146.16/.76 and 146.52. For more information: Richard Ferguson, KAØDXM, 1150 Albion Rd., Boulder, CO 80303. (303) 499-2871.

GEORGIA: The Atlanta Hamfestival 1984, sponsored by the Atlanta Radio Club, June 16 and 17, at the Atlanta Civic Center. 70,000 square feet of air-conditioned exhibitor space and over 800 outdoor flea market spaces will be available. Flea Market $12.50 per space in advance; $15.00 at the gate for both days. Hamfest registration $5.00 in advance, $6.00 at the door. To be pre-registered for the Flea Market or Hamfest, we must receive your application and check by June 8. Pre-registration applications received after June 8 will be returned. Hours 8 AM to 5 PM on Saturday. 8 AM to 2:30 PM on Sunday. Talk in on 3.97 MHz, 146.22/82 and 146.94 simplex. For pre-registration or other information write Atlanta Radio Club, PO Box 77171, Atlanta, GA 30357.

IDAHO: Kootenai Amateur Radio Society presents Hamfest '84 at the North Idaho Fairgrounds, Ceur D'Alene, June 9, 8 AM to 4 PM. Swap tables available at no charge. RV's are welcome but no hookups available at the site. Come early for our Friday program including pot luck and dancing afterwards. For further information write Avon Anderson, WB7WBZ.

ILLINOIS: The Six Meter Club of Chicago will hold its 27th annual Hamfest, Sunday, June 10, Santa Fe park, 91st and Wolf Road, Willow Springs. Gates open 6 AM. Advance registration $2.00, $3.00 at gate. Large swapper's row, picnicking, displays in the pavillion, plenty of parking, refreshments, AFMARS meeting and more. For advance tickets: Val Hellwig, K9ZWV, 3420 South 60th Court, Cicero, IL 60650 (or any club member). Talk in K9ONA 146.52 or K9ONA/R 37-97.

INDIANA: The Wabash Valley Amateur Radio Association's 38th annual Hamfest, Sunday, June 3, Vigo County Fairgrounds, Terre Haute. For more information SASE to WVARA, PO Box 81, Terre Haute, IN 47808.

INDIANA: The 16th annual Wabash County Hamfest, May 20, 4-H Fairgrounds, Wabash. 6 AM to 4 PM. Contact Don Spangler, W9HNO, 235 Southwood Drive, Wabash, IN 46992. (219) 563-5564.

INDIANA: The Annual Evansville TARS Hamfest, May 20, all indoors at the Vanderburgh County 4-H Fairgrounds. Open 6 AM CDT. Admission $3.00. Indoor tables $7.50. Outdoor flea market $3.00. Talk in on 147.75/.15 and 146.19/.79. For table reservations and information contact Mike Anderson, KA9LQM, PO Box 3284, Evansville, IN 47732.

MARYLAND: The Maryland FM Association's annual Hamfest, Sunday, May 27, Howard County Fairgrounds, West Friendship, 30 miles west of Baltimore. 8 AM to 4 PM. Donation $3.00. Tailgating $3.00. Inside tables $6.00 each in advance, $10.00 at the door if available. Talk in on 146.16/76 and 146.52. For table reservations or information: MFMA Hamfest Committee, c/o John Elgin, WA3MNN, 8216 Styers Ct., Laurel, MD 20707. (301) 621-2352.

MICHIGAN: The Chelsea Swap and Shop, Sunday, June 3, 8 AM to 2 PM, Chelsea Fairgrounds, Chelsea. Setups 5 AM. Donation $2.50 advance and $3.00 at the gate. Children under 12 and non-ham spouses admitted free. Talk in on 146.52 simplex and 147.855 Chelsea Repeater. For information: William Altenberndt, 3132 Timberline, Jackson, MI 49201.

MINNESOTA: The North Area Repeater Association will sponsor the state's largest Swapfest and Exposition for Amateur Radio operators, Saturday, June 2, Minnesota State Fairgrounds, St. Paul. Admission $4.00. Exhibits, booths, giant outdoor flea market. Free overnight parking for self-contained campers on June 1. Call wide area repeaters 25/85 or 16/76 for directions. For information write Amateur Fair, PO Box 857, Hopkins, MN 55343. (612) 420-6000.

NEW ENGLAND: The Hosstraders Spring Tailgate Swapfest, Saturday, May 12, sunrise to sunset at Deerfield, NH, Fair-

grounds. Admission $2 includes tailgaters and commercial. Friday night camping for self-contained rigs at nominal fee. None admitted before 4 PM Friday. No reserved spaces. Profits benefit Boston Burns Unit of Shriners' Hospital. Last year's donation over $4700.00. For map to northeast's biggest Ham Flea Market SASE to Norm, WA1IVB, RFD Box 57, West Baldwin, ME 04091 or Joe, K1RQG, Star Route Box 56, Bucksport, ME 04416 or Bob, W1GWU, Walton Road, Seabrook, NH 03874.

NEW HAMPSHIRE: The 10th annual Eastern VHF/UHF Conference, May 4-6, Sheraton Tara, Exit 1, US 3, Nashua. Friday night hospitality room. Tech talks by well-known VHFers. Pre-registration $14.50 to Rick Commo, K1LOG, 3 Pryor Rd., Natick, MA 01760 before April 29. Registration at door $20.00. Saturday night banquet is $15.00 payable before April 29. For information: Lewis D. Collins, W1GXT, 10 Marshall Terrace, Wayland, MA 01778. (617) 358-2854 before 10 PM.

NEW JERSEY: The Jersey Shore Chaverim is sponsoring the third annual Ham & Computerfest, June 10, 9 AM to 4 PM, Jewish Community Center, 100 Grant Avenue, Deal. 7300 sq. ft. of indoor space. Admission $3 per person (children under 12 and XYL's free). Refreshments available. Indoor table $8 and tailgating $3.50 per space. For reserved space SASE with advance payment to Jersey Shore Hamfest, PO Box 192, West Long Branch, NJ 07764 by June 1. Talk in on 147.045 +.6, 145.110 +.6 and 146.52 simplex. Deal, NJ is less than 50 miles from NYC and 70 miles from Philadelphia. For information call Arnold, W2GDS (201) 222-3009.

NEW JERSEY: TCRA Hamfest Tri County Radio Association, rain or shine, Sunday, May 13, Passaic Valley Community Center off Valley Road, Stirling, NJ. 9 AM to 4 PM. Indoors, refreshments, rest rooms, free parking. Tables $6, registration $2.50. Table reservations call or write Dick Franklin, W2EUF (201) 232-5955 or 270-3193, PO Box 182, Westfield, NJ 07090.

NEW YORK: "ROME HAM FAMILY DAY", Sunday, June 3, Beck's Grove in Rome. Presented by the Rome Radio Club, Inc. this event features something for everyone. Games, contests and the largest Flea Market in the area. Good food and beverages available throughout the day. Educational and scientific displays and presentations climaxed by a fine dinner and our "Ham of the Year" award. For further information: Rome Radio Club, PO Box 721, Rome, NY 13440.

NEW YORK: The 25th annual Southern Tier Amateur Radio Club's Hamfest, Saturday, May 5, Treadway Inn, Owego. Flea market opens at 8 AM. Vendor displays and sales. Tech and non-tech talks. Refreshments available. Dinner is at 6:30 PM by advance tickets only. Talk in on 22/82, 16/76, or 146.52 simplex. For further information SASE to KF2X, C. England, RD #1, Box 144, Vestal, NY 13850.

OHIO: The Sandusky County and Ottawa County Combined Hamfest, May 20, Ottawa County Fairgrounds, state Rt. 163, 3 miles east of Oak Harbor. Advance tickets $2.50, $3.00 at gate. Free trunk space and parking. Tables available. For information: John Dickey, 545 N. Jackson St., Fremont, OH 43420. Talk in on 147.675/075 or 52 simplex.

OHIO: The Champaign-Logan Amateur Radio Club's annual Hamfest, Sunday, June 10, Logan County Fairgrounds, Bellefontaine. Gates open 8 AM. Tickets $2.00 advance, $2.50 at the door. Tables $3.00. Plenty of free parking. Call in 147.60/00. Mobile check 146.52 simplex. For ticket info: Steve Kidder, N8ETD, Box 265, Russells Point, OH 43348 or (513) 843-6099.

OKLAHOMA: The Broken Arrow ARC and the Tulsa ARC will sponsor the Greencountry Hamfest, May 18, 19, and 20, Western Hills Lodge, 6 miles east of Wagoner at Sequoyah State Park. Pre-registration $2.50 or $3.00 at the door. There will be programs for the whole family. For information: Broken Arrow ARC, PO Box 552, Broken Arrow, OK 74012.

OHIO: Medina County Hamfest, sponsored by the Medina Two Meter Group, May 13, Medina County Community Center Building, Lafayette Rd., State Rt. 42 S.W. 8 AM to 4 PM. Vendor setup 7 AM. Refreshments and free parking. Tickets $2.50 advance, $3.00 at door. Tables $5.00. Some elec. hookups available. Talk in on 147.63/03, K8TV/R. For tables and tickets write PO Box 452, Medina, OH 44258. (216) 725-5021 or (216) 723-5010.

OKLAHOMA: The Great Plains ARC's third annual Northwest Oklahoma Eyeball & Swapmeet, Sunday, May 20, starting at 9 AM in Mooreland. Covered dish dinner at noon. Local airport. Dealer and swap tables free. Talk in on 147.72/12 and 146.52 simplex. Campsites available. For further information call (405) 994-5394 or write KB5XI, Gordon Richmond, Rt 1, Box 12, Mooreland, OK 73852.

PENNSYLVANIA: The Murgas ARC (K3YTL) will sponsor the annual Wilkes-Barre Hamfest, Sunday, June 3, 109th Armory, Market St., Kingston (across the river from Wilkes-Barre). General admission 8 AM. Setup at 6 AM. Admission $3.00. XYL's and children under 16 free. Tailgating $2.00 per space indoor/outdoor rain or shine. Talk in on 146.01/61 and .52 simplex. For further information: Hamfest Committee, PO Box 1094, Wilkes-Barre, PA 18703.

PENNSYLVANIA: The tenth annual Warminster Amateur Radio Club's Hamfest, Sunday, May 20, Middletown Grange Fairgrounds, Penns Park Road, Wrightstown, rain or shine. Gates open 7 AM (Vendors at 6 AM). Donation $3.00. Pre-registration $2.00. XYL's and children free. Tailgaters $2.00 add. per 10' space. Food and drink available. Talk in on 147.69/09 or 146.52. For information and pre-registration contact: Bill Cusick, W3GJC, Apt. 706 - Garner House, Hatboro, PA 19040. (215) 441-8048.

PENNSYLVANIA: The 2nd annual Southern Alleghenies Hamfest, May 13, 8 AM to 5 PM, Bedford County Fairgrounds one mile west of Bedford on Rt. 30 and 1/2 mile west of Rt. 220 bypass. Sponsored by the Bedford, Altoona, Somerset, PA, and Cumberland, MD ARCs and Blue Knob Repeater Association. Admission $3.00. Inside spaces $5.00 each, outside tailgating $2.00. Visit nearby restored Old Bedford Village at special Hamfest rates. Talk in on Bedford repeater 145.49 and 146.52 simplex. For more information call Tom Gutshall, W3BZN (814) 942-7334.

PENNSYLVANIA: The tenth annual Northwestern Pennsylvania Hamfest, May 5, Crawford County Fairgrounds, Meadville. Gates open 8 AM. Bring your own tables. $5 per table to display inside. $2 per car space outside. $3 admission, children under 12 free. Refreshments. Commercial displays welcome. Talk in on 145.13, 147.21, 147.03. Details: C.A.R.S., PO Box 653, Meadville, PA 16335. Att: Hamfest Committee.

PENNSYLVANIA: The 30th annual Breeze Shooters' Hamfest, Sunday, June 3, 9 AM to 4 PM, White Swan Amusement Park, PA. Rt. 60 near greater Pittsburgh International Airport. Free flea market and free admission. Family amusement park, food on site. Registration $2.00 or 3/$5.00. Talk in on 28/88 or 29 MHz. For information: Don Myslewski, K3CHD, 359 McMahon Road, North Huntingdon, PA 15642. (412) 863-0570.

TENNESSEE: The Radio Amateur Club of Knox County will hold its 18th annual Hamfest, May 26 and 27, Kerbella Temple Auditorium, Knoxville. Saturday 9 to 5. Sunday 10 to 4. Admission $3.00. Radio and computer forums, dealers, indoor/outdor flea market. Free parking. Talk in on 147.90/30. For information: Larry Poore, N4EHR, 4320 Felty Drive, Knoxville, TN 37918. (615) 687-3154.

VIRGINIA: The Tenth annual Manassas Hamfest, Sunday, June 3, Prince William County Fairgrounds, VA. RT. 234, 1/2 mile south of Manassas. 7 AM tailgate setup. 8 AM general admission. Tailgating, indoor exhibits, food available on grounds, YL program, CW proficiency awards. Admission $4 per person, under 12 free. Contact: Bob Kelly, KA4NES, General Chairman, Manassas Hamfest, c/o Ole Virginia Hams ARC, Inc., Manassas, VA 22110. (703) 361-9468.

WASHINGTON: The Yakima ARC's Central Washington State Hamfest, May 12 and 13, Hobby Building, Central Washington State Fairgrounds, Yakima. Saturday from 9 AM to 5 PM with lunch available. Sunday from 8 AM to 2 PM, breakfast and lunch available. Registration $4.00 advance, $5.00 at the door. Dealers' displays and a free swap and shop with plenty of tables. Talk in on 146.01/61 and 146.34/94. For pre-registration: Bob Rutherford, PO Box 9211, Yakima, WA 98909.

THE 574TH AND 565TH S.A.W. BNS. will hold their second reunion July 1984 in St. Louis, MO. Former members please write to Chas. A. McGaffin, San Mateo Rd., San Mateo, FL 32088. Phone (904) 328-9576 or to Angel M. Zaragoza, W6ZPR, 1571 - 9th St., San Bernardino, CA 92411. Phone (714) 889-2380 for full details.

OPERATING EVENTS

"Things to do..."

MAY 11 AND 12: HANDI-HAM System's special events station, WØEOO, will operate from Camp Courage, Maple Lake, MN, during the System's 15th annual Spring Convocation. For a special certificate SASE to: Handi-Hams, 3915 Golden Valley Rd., Golden Valley, MN 55422.

MAY 12 AND 13: ARMED FORCES DAY AT WEST POINT. The Meadowlands ARA will operate at the U.S. Military Academy in honor of Armed Forces Day 1964 using the club station call N2BMN. To confirm QSO send large SASE (8½ x 11) with 37¢ U.S. postage to POB 324, Little Ferry, NJ 07643.

MAY 19: In observance of Armed Forces Day, the U.S. Air Force Museum at Wright-Patterson Air Force Base, will host the operation of an Amateur Radio special event station. Listen for K8DMZ from 1400Z to 2200Z. To commemorate the event, the Museum will issue a special certificate for each two-way contact.

MAY 19: In recognition of the 35th annual Armed Forces Day celebration, Amateur Radio Station W4ODR, located aboard Naval Air Station Memphis, Millington, Tennessee, will be operating from 1400Z to 2200Z. Special certificates and QSL cards will be available to those who work W4ODR. QSL to ARS W4ODR, PO Box 54278, Millington, TN 38054.

MAY 19: ARMED FORCES DAY military-to-Amateur cross band operations will be conducted from 19/1300 UTC to 20/0245 UTC May 1984. East coast stations commence operations at 19/1300 UTC and west coast stations commence operations at 19/1800 UTC May 1984. Military stations will transmit on selected military frequencies. The military operator will announce the specific Amateur band frequency being monitored. Entries must be postmarked no later than 26 May 1984 and submitted to the respective military commands. Stations copying AIR send entries to: Armed Forces Day Test, 2045CG/DONJM, Andrews AFB, DC 20331. NAM, NAV or NPG to: Armed Forces Day Test, HQ Navy-Marine Corps MARS, 4401 Massachusetts Ave., N.W., Washington, DC 20390. WAR to: Armed Forces Day Test, Commander 7th Signal Command, Att: CCN-PO-OX, Fort Ritchie, MD 21719.

MAY 25 AND 26: ROYAL CANADIAN Air Force (RCAF) Telecommunications reunion. To honor the 50th anniversary of Air Force communications the reunion will be held at the Canadian Forces School of Communications and Electronics at Kingston, Ontario, for all active duty and retired members and spouses. For more information write to Air Force Telecom Reunion Committee, CFB Kingston, Kingston, Ontario K7L 2Z2.

MAY 26: NISKA-DAY '84. The Niskayuna, NY, High School Club Station, WB2OKK, (OK Kids) will operate from 1500Z to 2100Z to commemorate the 175th anniversary of the community of Niskayune. For a commemorative QSL card SASE to ARS WB2OKK, Niskayuna High School, 1626 Balltown Road, Niskayuna, NY 12309.

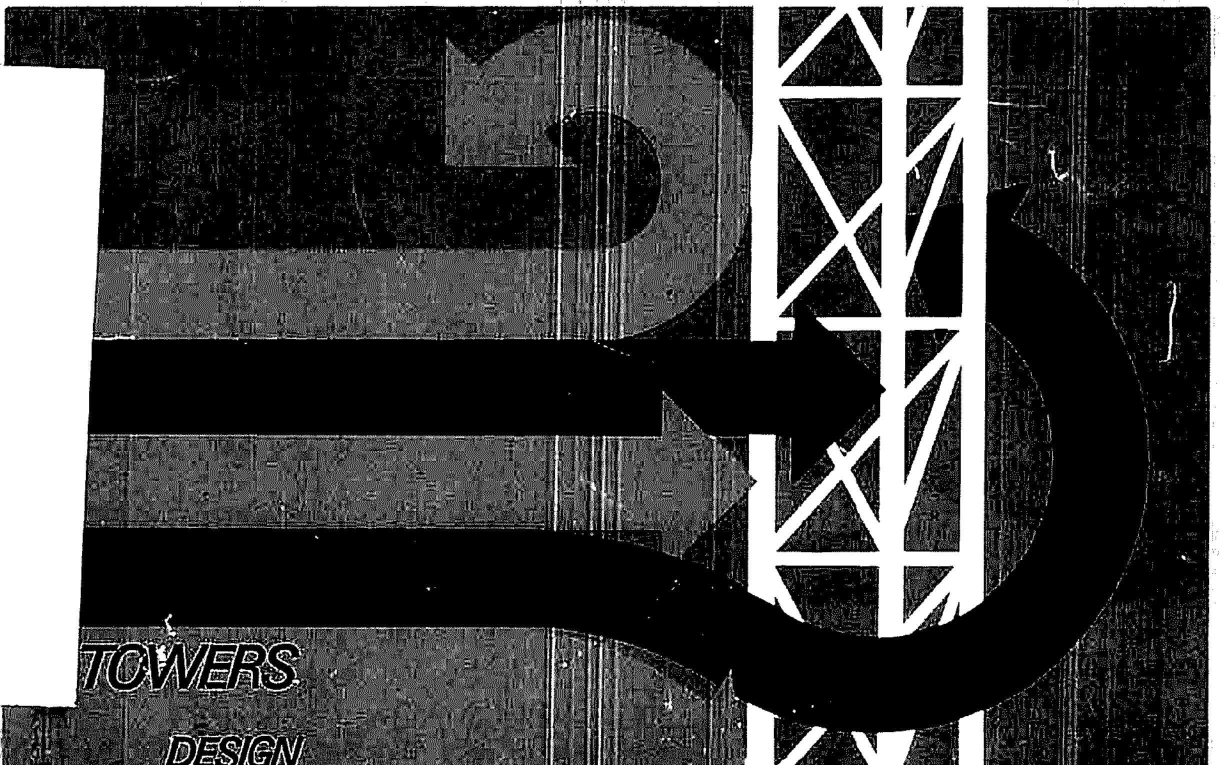
JUNE 2 AND 3: W.I.N.O., the Wireless Institute of Northern Ohio, an organization sponsored by the Lake County ARA, will operate a special events station to commemorate Ohio Wine Week. Listen for K080 operating from an actual winery in Madison, Ohio. A special 8½ x 11 certificate will be available from: K060, WINO Weekend, 7126 Andover Drive, Mentor, Ohio 44060.

JUNE 8 AND 9: Madison County ARC will operate club station W9VCF, portable from the historic Eight Street Festival in Anderson, Indiana. A special certificate will be offered to persons contacting the club station during the festival or any club member during the month of June. Send log info and $1.00 donation to Madison County ARC, c/o Frank M. Dick, WA9JWL, 921 Isabelle Drive, Anderson, IN 46013.

JUNE 9 AND 10: The Knox County ARC will operate a special event station to commemorate Galesburg Railroad Days, an annual event for Galesburg, Illinois. Listen for W9GFD. For a special commemorative QSL card SASE to Knox County ARC, W9GFD, 1694 Bluebird Drive, Galesburg, IL 61401.

JUNE 1984 / $2.50

ham radio magazine



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also: the peaked lowpass — a look at the ultraspherical filter • microphone calibration • impedance matching • applied Yagi antenna design • VHF/UHF world • ham radio techniques

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JUNE 1984

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contents

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REFLECTIONS

SCORE: 99 FOR . . . 1 AGAINST

Of the thousands of towers installed in the United States last year, 99 percent remain standing. This is truly a remarkable statistic, considering how many of these structures were installed using less than the best techniques and, to boot, were hardly ever maintained.

Of those that came down, we could say that some unusual weather condition (hurricane, tornado, or ice storm) or some other extraordinary occurrence was to blame — but stop and think how many of these would still be standing today if the proper foundation had been laid, if wire instead of nylon rope had been used for guys, if that 10 feet of additional masting hadn't been used .

Do you recognize any of these situations? Is your present installation ready to become a negative statistic? If so, read on.

O.K., we're all Amateurs, but just for the moment, let's borrow some good advice from the engineering disciplines (or maybe it's just good old common sense). Before putting up that 6-element super deluxe 27 dBn gain (decibels with respect to a conductive noodle) Yagi atop the 102-foot perch, sit down, take pencil in hand and think it through. A safe tower/antenna installation (or any installation for that matter) can and should be engineered in three simple steps: DESIGN it according to the manufacturer's specifications, INSTALL it correctly, and MAINTAIN it. Unfortunately, most of us (and I don't exclude myself) are in such a rush to become operational — taking advantage of the good weather and time off from work, vacation, or chores — that we just can't afford to take the time to do it right. Or can we? Can we afford to come home one day and find an irate neighbor shouting and pointing to our aluminum thing of beauty, still attached to our tower, protruding upside down from *his* roof?

Just because we're Radio "Amateurs," it doesn't mean that our installations have to be "amateur." And it doesn't mean that all commercial installations are necessarily safer. A case in point: a friend of mine was doing a site survey for a cable TV outfit and had to climb one of the existing towers for the preliminary observation phase. Well, it was getting late in the day and, to be quite honest, my friend was a wee bit leery about climbing that tower because of something unnerving — but not readily identifiable — about the foundation and the sandy soil. When he came back the next day to climb the tower, he found the tower on the ground, having fallen by its own accord during the night.

To help prevent this kind of near-disaster, and real disasters as well, the theme of this issue (if you haven't already guessed) is *towers*. In a sequence of articles not dissimilar to a Greek trilogy, Roger Cox (WB0DGF), John Haerle (WB5IIR), and Steve Makulec (KB9IW) discuss the three important steps of tower construction: design, installation, and maintainance — and there's nothing mythological about that.

Rich Rosen, K2RR
Editor-in-Chief

presstop de W9JUV

PRESSURE IS BUILDING IN THE 900 MHZ SPECTRUM for a variety of possible uses, at least some potentially in conflict with the Amateur 902-928 MHz allocation that came out of the 1979 WARC. Mura, a leading producer of cordless phones and other consumer electronics, has petitioned the FCC to establish a new "Consumer Radio Service" at 935-938 MHz. Mura's proposal would provide 120 channels using automatically identified radios with a five-mile range at a projected cost of under $100 each. Mura also proposes phasing out 27 MHz and GMRS CB, and rejects GE's earlier proposal for a 900 MHz service with repeaters and telephone interconnect as simply a variation of the new cellular communications system.

Cordless Phones Are Yet Another Prospective 900 MHz User, proposed by the Electronic Industries Association in another Petition for Rule Making. The EIA wants two 2-MHz bands in the 900 MHz region, separated by at least 45 MHz, to become available by January 1, 1987. The FCC did recently add 10 new "interim" cordless phone channels at 46/49 MHz.

All This Is In Addition To The Rapid Growth Of Paging, cellular radio and other two-way services in the 800-1000 MHz spectrum. However, on March 28 the ARRL filed its own Petition for Rule Making, asking for early implementation of privileges for Technician and higher class licensees on the 902-928 MHz Amateur band. The FCC reports very little previous Amateur input about this band. At the same time, the ARRL also asked for use of the new 24-MHz band, by General Class and higher. Yet another petition, by five Amateurs who all hold experimental licenses for both the new 18 and 24-MHz Amateur allocations, asks that Amateur operation be permitted on both bands on a non-interference basis.

THE FIRST VOLUNTEER-ADMINISTERED AMATEUR EXAMS WERE GIVEN APRIL 17 IN ALASKA, giving the No. 1 VEC, Anchorage Amateur Radio Club, still another first in the program. 42 Tech and General exams were given in their session. The Dayton ARA wasn't far behind, giving about 400 exams during its April 28-29 Hamvention weekend. Dissatisfaction over ARRL inaction on the VEC program seems widespread, and rumors of some sort of forthcoming shakeups at Newington were circulating at the Hamvention.

Amateur Frustration With Lack Of Upgrade Opportunity is now becoming serious in some areas. The third call area, for one, is now booked solid through the year's end, when FCC-administered exams end and the volunteer program must take over. However, a third call area VEC, the Laurel (Maryland) ARC, is expected to be certified shortly.

A Seventh District VEC, The Boeing Employees Amateur Radio Society, has been accepted by the FCC, and the Dallas ARC has been certified as the VEC for the fifth district. In the second and eighth call areas additional VECs are also coming on board. However, the first, zero, and sixth districts still have no known prospects, though a sixth district group is reported to be preparing to make a proposal.

The FCC's NPRM On Exam Fees Drew Only a Handful Of Comments from several clubs, a few concerned individuals, and of course the ARRL. Most supported the concept of expense reimbursement for examiners as well as the VECs, under the direction of the VECs. It is hoped the Commission will be able to authorize fee collection before the summer recess.

The Revised Novice Exam Program, In Place Since Last Summer, is working well for the most part, though Gettysburg says they still have some problems. Some Amateur examiners still don't realize they're supposed to write, administer, and grade the exams themselves, and are still writing Gettysburg for exams or sending applicants' answer sheets there for grading. The other major problem is with improperly filled out Form 610s: the examiner either neglects to certify that both the CW and written exams were passed, or neglects to include a signed statement certifying that the examination was properly administered.

OPERATING AFTER HIS LICENSE WAS LIFTED BROUGHT A SUSPENDED sentence with a threat of prison to a California ex-Amateur. The former N6BII, who lost his license for jamming WESCARS and other 40-meter operations was sentenced to a 90-day suspended sentence and three years probation in Federal District Court April 19. Under the terms of his probation, however, he can go to jail if he even talks over another Amateur's station during the probation period, unless the FCC chooses to relicense him.

EXPANSION OF ADDITIONAL PHONE BANDS IS STILL in the FCC's hopper, but staff cuts at the Commission and implementation of the Volunteer Exam Program have kept it on the back burner. There's hope that the Commissioners will be able to act on expansion before the summer recess, but what direction expansion (if any) will take is very much up in the air.

DAYTON HAMVENTION'S "HAM OF THE YEAR" IS DAVE BELL, W6AQ, honored "...for dedicated use of his Cinematographic skills to bring the story of Amateur Radio to the world." This year's "Special Achievement Award" went to Ethel Smith, K4LMB, for her efforts on behalf of the YLRL. The "Technical Achievement Award" went to Lyle Johnson, WA7GXD, for development of the Tucson Terminal Node Controller for packet radio. Congratulations to all!

A NUMBER OF SCHOLARSHIPS FOR AMATEURS who are now or will be attending a full-time college are being offered by the Foundation for Amateur Radio. Write them at 6903 Rhode Island Ave., College Park, Maryland 20740, for application information.

EXTENSIVE REVISION OF 2-METER REPEATER FREQUENCIES ALONG THE U.S.-MEXICAN border may be coming in the wake of recent interference problems. Mexican government authorities have met with their Amateurs to discuss on-going across-the-border coordination difficulties, and have proposed joint meetings with the FCC and U.S. Amateurs to resolve them.

comments

can you help?

Dear HR:

I need a schematic for the ES 221K frequency counter advertised by ESE in the October, 1974, issue of *ham radio*, page 90. I have a problem with the counter, but no manual or schematic.

A.J. Massa, W5VSR
New Orleans, Louisiana

Dear HR:

I need a schematic and/or manual for a Model 555/N oscilloscope made by the Data Instruments Division of Pennsauken, New Jersey. Upon writing to them I found that they have moved (or closed) without leaving a forwarding address.

I am willing to pay for a photocopy of this data. Please contact me at the address below.

Tom Adams, K9TA
110 N. Few St. Apt. #2
Madison, Wisconsin 53703

aiming for excellence

Dear HR:

Your excellent editorial ("Reflections," February, 1984, page 6) really tells it like it is. I wish every Amateur would read it and take your advice.

In most endeavors people strive for excellence. The shooter, golfer, bowler, artist, craftsman, and so on, aspire to excellence — why shouldn't Amateur Radio operators seek a similar goal?

I made some friendly suggestions recently to a group of several newly licensed Amateurs on how to improve their operating procedure. You would not believe the reaction to my suggestions. There were such comments as:

"You have no responsibility for the way we choose to operate."

"Ham radio is a hobby for the average citizen and was not intended for engineers — hence the name, *Amateur* Radio."

"You old-timers want to take over and run us newcomers out."

"You give technical talks to the club only to show how smart you are."

These are pretty sick comments, but of course, they come from a very few insecure people. The majority of hams are mighty fine people.

I operated a DX station for three years in China and the Phillipines. When I encountered the operating rudeness to which you refer I had a very effective response. Even though the signal was very loud, I would report it as impossible to copy or even ascertain the call. A call on the QSO frequency was an immediate entry to my blacklist. Let me assure you that some pretty big names in the DX ratrace never got a QSO from me while I was in Canton, Swatow, Amoy, or Foochow.

Keep up the good work.

I.L. McNally, K6WX
Sun City, California

better SSB

Dear HR:

I was pleased to read the article: "Better-Sounding SSB" in the February, 1984, issue of *ham radio*. I for one have always maintained that SSB should sound just as good as the voice at the mike! However, for years "communications quality" has been in vogue. The simple changes mentioned by Mr. Measures are good. Now if some enlightened manufacturer would just lead the way and recognize that many hams would prefer smooth, undistorted audio, I might finally buy a new rig!

Let's have more such articles!

James E. Taylor, W2OZH
Webster, New York

optical receiver cost

Dear HR:

I found Poon and Pieper's "Construct an Optical FM Receiver," (November, 1983, page 53) very interesting. The drawings and equations resemble those found in my engineering log book as of late. This is quite state-of-the-art technology, and experimentation should be encouraged; however, the authors did not mention the cost of their project. To prevent anyone seriously interested in persuing this project from becoming discouraged, I thought I might pass along my costing estimate of the items specified. Depending on the type and power of laser purchased and quality of lenses used, the cost of the project will run between $858.00 and $1228.00. (This, by the way, is relatively cheap for a basic acousto-optic receiver.)

Good luck.

David A. Clingerman, W6OAL
Newbury Park, California

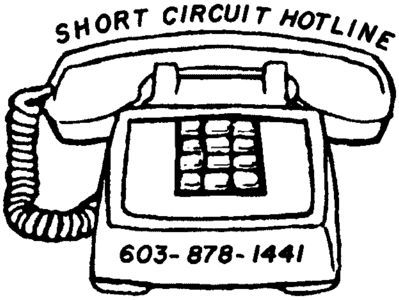
(A call to Edmund Scientific revealed that an Aerotech 1 milliwatt Helium-Neon laser could be purchased for $330 and a complete set of spherical/cylindrical lenses for under $50 . . . **Ed**)

HF operations

Dear HR:

I would like to correspond with anyone involved in HF mobile operation or the design of HF solid-state amplifiers.

Phil Zelter-Jenkins
70, Cross Oak Road, Berkhamsted
Hertfordshire, HP4-3HZ, England



TOWER DESIGN:

match your antenna to your tower

Here's how to calculate wind area, wind load, and bending moment

Have you heard the story *about the ham across town who bought a new $500 long-boom tribander to replace his older, smaller beam? He rented a "cherry-picker" at $50 an hour to have it installed on top of the tower he'd originally bought for his 5/8-wave CB antenna in the days before he'd become a ham. The rotator, a small CDE unit he'd bought for $10 at a flea market, had worked fine with the TH3JR that he'd also purchased soon after getting his ham ticket.*

A few months later, a snowstorm driven by 50 MPH winds demolished his entire antenna system, and worse yet, sent the tribander crashing through the roof of his neighbor's house. He wound up in a hassle with his insurance company over the damages, and the neighbor soon began circulating a petition to ban antennas and towers from that city.

You don't think this could happen? Look at the towns of Burbank, Illinois, and Cerritos, California, and see what Amateurs in those communities are fighting. A ruined antenna and tower may not only rob your wallet, but may also rob you and your friends of your operating privileges because of the negative publicity that inevitably follows such an incident.

the match game

When someone says that an antenna is perfectly "matched," most people assume that he or she means that the VSWR is a perfect 1:1. But there's another kind of "match" that's at least as important as VSWR and essential in preventing the kind of mayhem described above. This is the match that occurs when an antenna is *physically* mated to its mast, rotator, tower, and any other antenna on the tower.

How can you determine whether your antenna is properly "matched"? Well, for starters, check the specifications of the equipment you're now using. Don't wait until something happens; you may have added a small VHF antenna or a 30/40 meter add-on kit without realizing that you've exceeded the wind-loading of your tower or *rotator.*

Start by referring to your antenna instruction or assembly manual. All legitimate antenna manufacturers include both electrical and mechanical specifications with their products. If you can't find your manual, or if the specifications are not listed, write or call the manufacturer's customer service department for this information.

wind area

Look for the entry titled "Wind Area" or "Effective Surface Area" under "Mechanical Specifications." This number, expressed in square feet or square meters, is a measure of the physical size of the antenna. It represents the maximum surface area against which the wind could theoretically push. The total wind area figure should represent the "worst-case" surface area of the antenna — a combination of both the total boom and total element surface areas.

If two or more antennas are mounted within approximately 2 feet of each other, their wind areas can simply be added together to provide a total Antenna System Wind Area. This wind area should be equivalent to or less than the rated wind area maximums of the rotator and tower.

As an example, consider a tribander with a wind area of 5.7 square feet (0.53 square meter) and a 2-meter vertical with an area of 1.5 square feet (0.14 square meter) mounted 2 feet (0.61 meter) above the tribander. The total Antenna System Wind Area of this system would be 7.2 square feet (0.67 square meter),

By Roger A. Cox, WB0DGF, Telex Communications, Inc./Hy-Gain Division, 8601 Northeast Highway 6, Lincoln, Nebraska 68505

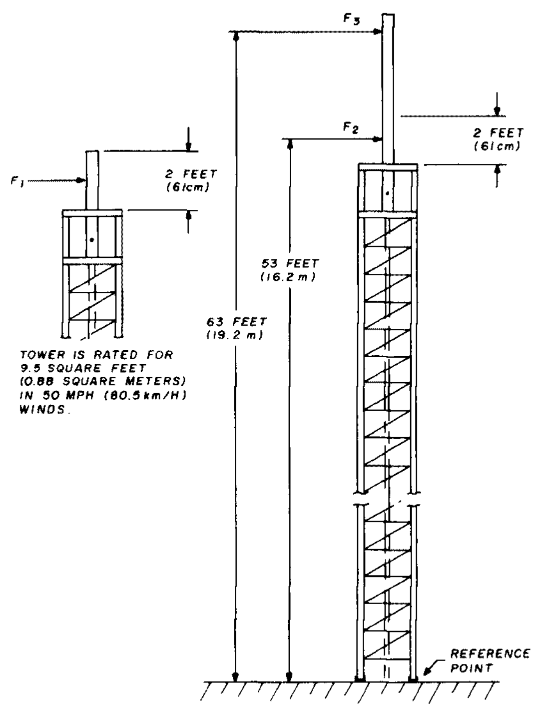


fig. 1. Locations of forces on tower masts.

and would easily match a tower rated at 9 square feet (0.84 square meter) or a rotator rated at 8.5 square feet (0.79 square meter).

Determining wind area becomes more complex when you want to stack two or more HF beams or numerous arrays of VHF Yagis, because under these circumstances the separation required for good electrical performance is almost always greater than 2 feet (0.61 meter).

To analyze complex antenna systems made up of antennas spaced *more* than 2 feet (0.61 meter) apart, you need to know how each antenna contributes to the loading of the entire system.[1] Check your manual under "Mechanical Specifications" once more; you should be able to find the "Wind Load" of the antenna at a specific wind velocity. [Usually a wind velocity of either 80 (129 km/hour) or 100 MPH (161 km/hour) is used.]

wind load

The wind load of an antenna is related to the wind area of the antenna through the equation:

$$F = PA \qquad (1)$$

where F is the wind load force in pounds, P is the wind pressure in lb./ft.², and A is the antenna wind area in ft.². P is dependent upon the wind velocity, V, and is usually* found from the equation:

$$P = 0.004\ V^2 \qquad (2)$$

where V is the wind velocity in MPH. At 80 MPH, $P = 25.6$; at 100 MPH, $P = 40\ lb./ft.^2$.

In **fig. 1** a tower with a wind load limit of 9.5 square feet in 50 MPH winds is shown. From **eq. 2** we can find the pressure, P, which the wind exerts on the antenna in a 50 MPH wind.

$$P = 0.004\ (50)^2$$
$$= 10\ lb./ft.^2$$

In order to find the maximum force which may be exerted on the tower, we use **eq. 1**.

$$F = PA = 10(9.5) = 95\ lbs.$$

Therefore the maximum allowable force within the 2-foot limit is 95 pounds.

the bending moment

In order to evaluate the effects of placing more than one antenna on a mast above the tower, we must look at the bending moment, M, with respect to the weakest point in our tower/mast system. The moment, M, is found by

$$M = FD \qquad (3)$$

where F is the force and D is the distance to the weakest point. The moments may be added together for the various antennas as long as the same point is used as a reference. In **example 1**, if we suspected that the tower base was our weakest point, and that our force F_1, was applied to the mast 53 feet above the point, the moment, M_1 could be expressed as:

$$M_1 = F_1D = 95(53) = 5035\ ft.\ lbs.$$

If we were to replace F_1 with an antenna that exhibited 50 pounds of force, and added another antenna 10 feet above it that exhibited 40 pounds of force, would we exceed the allowable moment of 5035 ft. lbs.? Your first guess may be no, because the forces add up to only 90 pounds and this is less than the 95 pounds we had before. And if you were to compare wind areas, you would find that the 50 pound antenna at 5 square feet and 40 pound antenna at 4 square feet also add up to less than 9.5 square feet, or rated wind area for this tower.

But if you were to add up the moments,

$$M = \Sigma\ FD \qquad (\Sigma = sum\ of\ldots)$$

$$M_2 = 50\ (53) = 2650$$

$$M_3 = 40\ (63) = 2520$$

$$M_2 + M_3 = 5170\ ft.\ lbs.$$

*The equation $P = 0.004\ V^2$ takes wind gusts and turbulence into consideration. For a steady (laminar) flow, the equation $P = 0.00256\ V^2$ should be used. Consult the Uniform Building Code (UBC) for a more thorough explanation.

you would find that they easily exceed the moment limit of 5035 ft. lbs.

The base of the tower may not always be the weakest point of your tower/mast system. Other susceptible points are the point of attachment of the top set of guy wires, a house bracket, a junction in a telescoping tower or mast, or even the point of attachment of the mast to the tower. If you've used short inserts to reinforce the mast, the end of any insert within the mast may be a point of vulnerability as well. Your best bet is to follow the tower and mast manufacturer's recommendations whenever you stack large antennas.

rotator wind area

As in determining the wind area of your antenna, you should begin analysis of your rotator's wind area by reading your instruction manual. If the specifications are not listed, or if you cannot find your manual, write or call the manufacturer's Customer Service Department for this information.

Look in the specifications section for the entry titled "Maximum Wind Area." There will be two different entries — one for mast mounting and one for inside tower installations. For mast mounting, there will also be a maximum distance of the antenna above the rotator provided; this is usually 2 feet (0.61 meter).

Just as in the previous example, if all of your antennas are mounted within this range (2 feet/0.61 meter) you may add the antenna wind areas together and compare that total to the rotator's rating.

loading

If you're using a small tower or mast, with the rotator installed on top, be extra careful to observe the rotator's mechanical limitations on side thrust. Use the procedures shown in the wind load section of this article with the rotator's rated wind area and wind speed to determine the maximum force that can be applied within 2 feet (0.61 meter) of the top of the rotator. If no wind speed is given, use a conservative figure such as 50 MPH.

As an example, consider the Hy-Gain CD4511. When mast-mounted, its rated wind area is 5.0 square feet (0.46 square meter). I will use 50 MPH as the wind speed. Using eqs. **1** and **2**:

$$F = 0.004\ V^2\ A = 0.004\ (50)^2(5) = 50\ lbs.$$

In this case, using a distance of 2 feet (0.61 meter) above the rotator, the moment is 100 ft. lbs. You can use the same procedures shown in the bending moment section to evaluate the effects of placing more than one antenna on a mast above a rotator.

Although stacking antennas above a rotator installed on top of a mast or tower is not recommended, it can be done if the maximum moment limitations are adhered to. The weakest point in this case is assumed to be the center of the rotator.

Although slightly exceeding the maximum ratings of a rotator may not break or permanently damage it, doing so will more than likely impair its operation in some way, especially during windy conditions. Continued use in this manner is sure to shorten the useful life of your rotator.

If you install a rotator inside a tower, a different set of mechanical limitations applies. The side thrust is now less important, but the stall and braking torque of the rotator is more important. This is why a different wind area rating is listed for rotators installed within a tower.

The rotator's wind area rating within a tower is usually related more to the braking power rather than to the stall torque of the rotator. The braking power is the maximum allowable torque that the antenna load may present to the rotator without causing it to rotate. This torque is usually not a steady torque, but rather a pulsing, almost sinusoidal torque produced by the antenna rocking back and forth in a violent wind storm.

moment of inertia

The amount of torque produced by this rocking action is directly related to the moment of inertia of the antenna. The moment of inertia is similar to the bending moment as discussed earlier, because it is related to force and its distance from a reference axis. However, complex structures such as antennas must be analyzed as the summation of all the moments produced by the various portions of the antenna. Also, since each moment is directly related to the mass of each antenna portion and the square of its distance to the axis, an antenna with a very long boom with heavy elements will have *more* total moment of inertia than an antenna with a short boom with very light elements, even though they may have the same total wind area! This has produced some difficulty in assigning wind area ratings for rotators. Luckily most commercially available Amateur antennas have short enough booms so that their wind areas and moments of inertia are closely related. However, on long-boom homebrew, commerical, or military-type antennas, one has to be extremely cautious when selecting a rotator when given only a wind area rating.

Although the wind area rating of a rotator is not wholly determined by the rotator stall torque, a higher stall torque is required for turning larger antennas. On a calm day, the act of rotating the antenna produces wind against each element and boom. This wind is a force that opposes the direction of rotation. If, at full speed, this force produces the amount of torque required to stall the rotator, the rotator will slow down until the wind force is less than that which produces a stall. Therefore, a rotator with a small stall torque will sometimes turn more slowly than one with a higher

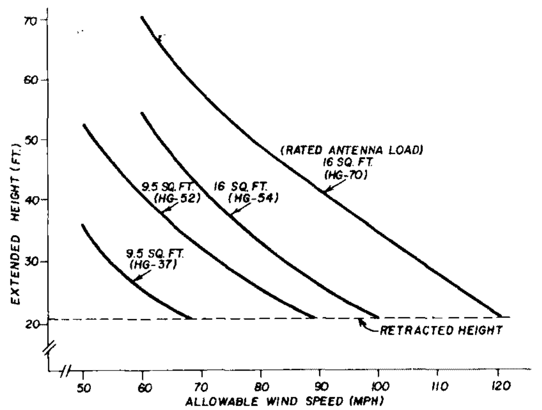


fig. 2. Extended height of self-supported crank-up tower is a function of wind speed and antenna location.

stall torque. On a windy day, the rotator with less stall torque may not be able to overcome the forces produced by the wind.

towers

Every tower ever commercially manufactured has some kind of instruction manual or specification guide, no matter whether it is a crank-up, guyed, or self-supporting tower built of steel or aluminum tubular or right-angle stock.

The manuals accompanying crank-up towers, with which I am most familiar, list a "wind load limit." This is the maximum wind area in square feet (or square meters) that the tower will *safely* hold at its maximum height at a particular wind velocity. The industry standard is to rate these towers at either 50 MPH (80.5 km/hour) or 60 MPH (96.6 km/hour).

Because of the specific nature of crank-up towers, the owner can crank the tower up and down for antenna installation and servicing and also crank the tower down whenever high winds are expected. With the tower completely retracted, the bending moment at the base of the tower caused by the antenna load is significantly reduced.

If you're willing to crank the tower down every time strong winds are expected, it's possible to expect your antenna system to survive near-hurricane conditions even with the maximum allowable antenna wind area. **Fig. 2** shows the wind velocities under which you can expect a Hy-Gain crank-up tower to survive given the maximum rated antenna loads at the extended heights shown in the graph.

For example, if you have a 9.5 square foot (0.8 square meter) antenna load on your Hy-Gain HG-52SS tower, which is cranked down to 21 feet (6.4 meters), you can expect your system to survive a wind velocity of 90 MPH (145 km/hour). (This assumes, of course, that the tower was properly installed, and that all the manufacturer's recommendations were followed.)

You can also expect your fully extended tower to survive higher wind velocities than specified if the antenna wind area is less than the maximum rating for the particular size tower. **Fig. 3** shows how the maximum antenna wind area for a given tower varies with the allowable wind velocity. For example, the HG-70HD, which is rated at 16 square feet (1.5 square meters) in 60 MPH (96.6 km/hour) winds, can safely handle only 10 square feet (0.9 square meter) in 70 MPH (113 km/hour) winds.

As you can also see from **fig.** 3, larger antenna loads may be possible if lower wind velocity figures are used. Unless your tower is sheltered from the wind, it would be dangerous to assume that the wind would always stay under 30 MPH (48 km/hour). These figures should be used only as a demonstration of what may be possible in your installation. Other factors such as the type of soil, ice loading, and wind-driven sand (sandstorms) may affect your particular installation. Again, your best bet is to follow the manufacturer's recommendations on anything questionable.

The manufacturers of guyed towers (such as the Rohn Model 25G and Model 45G) usually recommend specific guying configurations depending on the height of the tower and the specific area of the country in which installation is planned.[3] The maximum wind area is specified as "allowable load" for each type of tower.

The Electronic Industries Association has divided the continental U.S. into 3 wind load zones (see **fig. 4**).[4]

Zone A encompasses most of the United States. Short towers constructed within this zone should be capable of withstanding loading of 30 pounds per square foot (147 kilograms per square meter). This corresponds to approximately 87 MPH (140 km/hour) winds.

Zone B encompasses northwest Washington, north-central California, part of the northern Great Plains and northern Rockies, the area surrounding Madison, Wisconsin, and most of the Gulf Coast and eastern seaboard. Short towers within this zone should be capable of withstanding loading of 40 pounds per square foot (196 kilograms per square meter). This corresponds to approximately 100 MPH (161 km/hour) winds.

Zone C encompasses two areas, the southeast tip of Florida and most of the eastern coast of North Carolina. Short towers within this zone should be capable of withstanding loading of 50 pounds per square foot (245 kilograms per square meter). This corresponds to approximately 112 MPH (180 km/hour) winds.

Fig. 5 shows typical guying configurations for the Rohn Model 25G guyed tower installed in Zone A at various heights. This configuration is for an "allowable load" of 6 square feet (0.56 square meter).

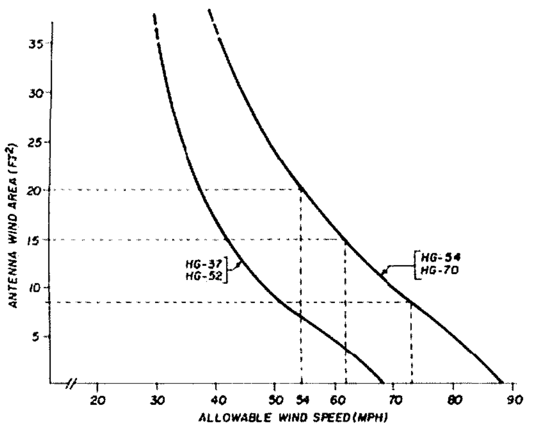


fig. 3. Example of antenna wind area versus wind speed for four commercially-available towers.

Self-supporting towers are similar to crank-up towers in their specifications. They usually specify a maximum wind area or wind load limit at a particular wind velocity. If a particular manufacturer does not list a wind velocity with the load rating, you should ask for this figure from its customer service department. If the wind velocity rating given is less than 50 MPH (80.5 kmph), you may wish to add guys to your tower if you are at or near the rated load. If the wind velocity given is greater than 80 MPH (129 kmph), the rated loading should be safe unless you live in Zones B or C, as previously mentioned. If you do live in one of these zones, be sure to choose a tower with a rated wind velocity figure greater than 100 MPH (161 kmph), or be prepared to add guys.

masts

While the mast doesn't receive as much attention as the tower or rotator in the ratings game, it can be of vital importance in maintaining the integrity of your antenna system.

Only a few manufacturers supply masts for antenna systems. Normally it's easier and cheaper for an individual to purchase a length of steel tubing at the local lumber yard or electrical supply store than it is to purchase it by mail order. For the average antenna installation, this is quite adequate. A length of 2-inch (51 mm) O.D. schedule 40 or schedule 80 pipe is suitable for a tribander and a small VHF antenna. However, if you plan to stack HF antennas in Christmas-tree fashion to assemble an array of VHF antennas for EME, you may wish to analyze your system and consider other possibilities.

Following the previous examples given, find the wind area for each antenna, boom, and mast in your system. Use these wind areas and an appropriate wind velocity to determine the loading from each. Multiply these loads by the distance to the nearest supporting boom, mast, or tower to find the bending moments of these points. To analyze the flexural strength at these points, you'll need to have information about the structural member at each point.

You will also need the initial moment of inertia, I, of the cross section about the neutral axis. This can be obtained from:

$$I = \frac{\pi(d_1^4 - d_2^4)}{64} \ in.^4 \qquad (4)$$

(for a circular cross section)
where d_1 is the members O.D. in inches, and d_2 is the members I.D. in inches.

U.S. WIND LOADING ZONES

Recommended wind loading zones values of wind loading in pounds per square foot for tower designs as recommended by Electronic Industries Association (RS222C).

WIND LOADS FOR VARIOUS AREAS IN U. S.			
Tower Height	ZONE A	ZONE B	ZONE C
Under 300 Ft.	30 Lbs/Sq. Ft.	40 Lbs/Sq. Ft.	50 Lbs/Sq. Ft.
301 Ft. to 650 Ft.	35 Lbs/Sq. Ft.	48 Lbs/Sq. Ft.	60 Lbs/Sq. Ft.
Over 650 Ft.	50 Lbs/Sq. Ft.	65 Lbs/Sq. Ft.	85 Lbs/Sq. Ft.
MAP CODE			

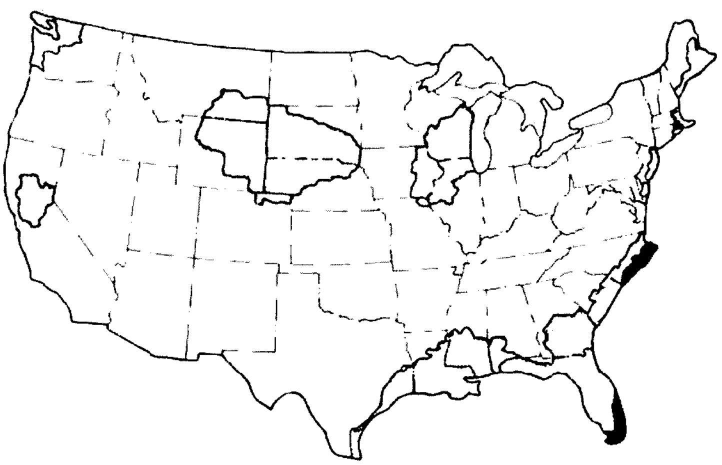


fig. 4. Chart details recommended wind loading on towers of various heights, per E.I.A. RS222C.

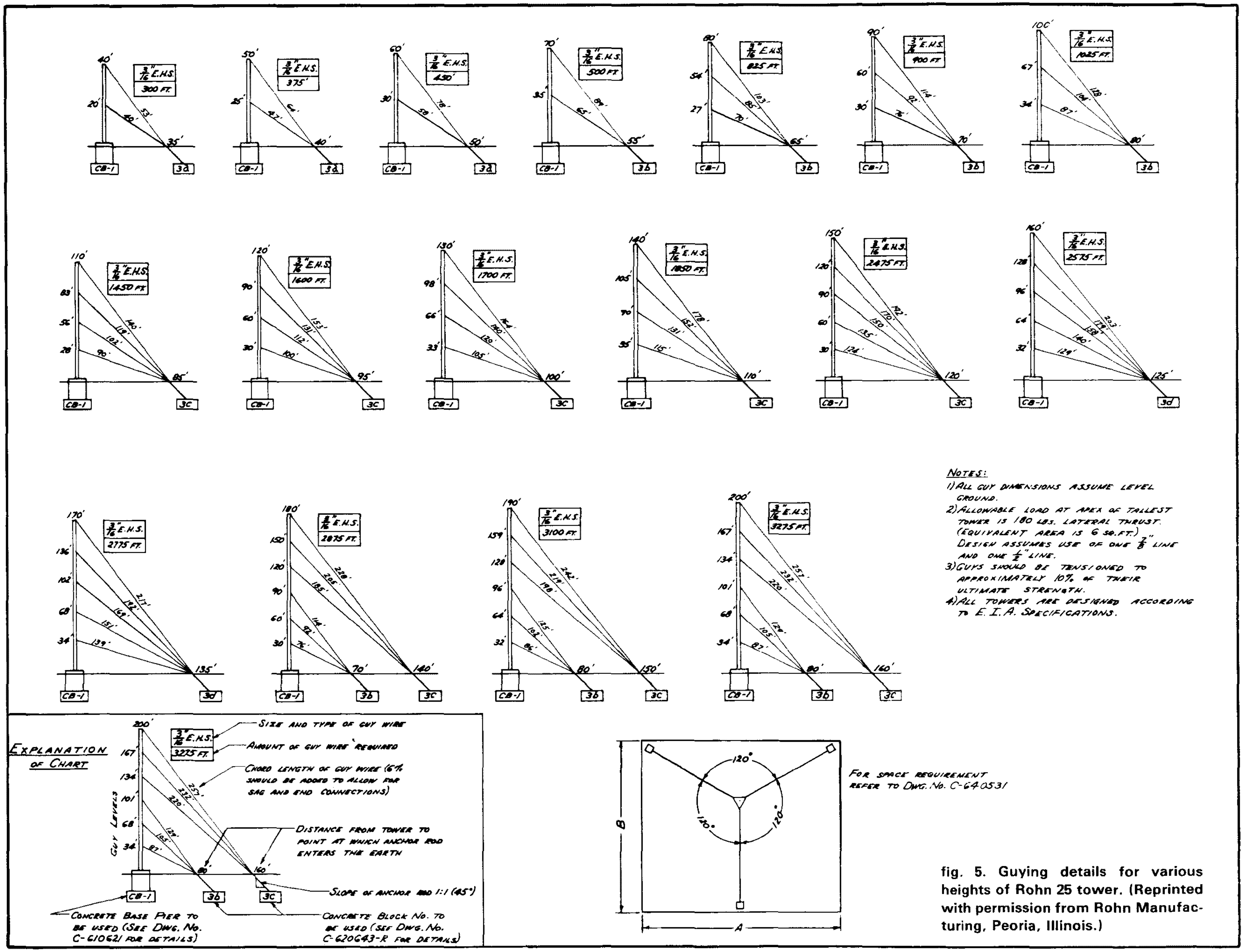
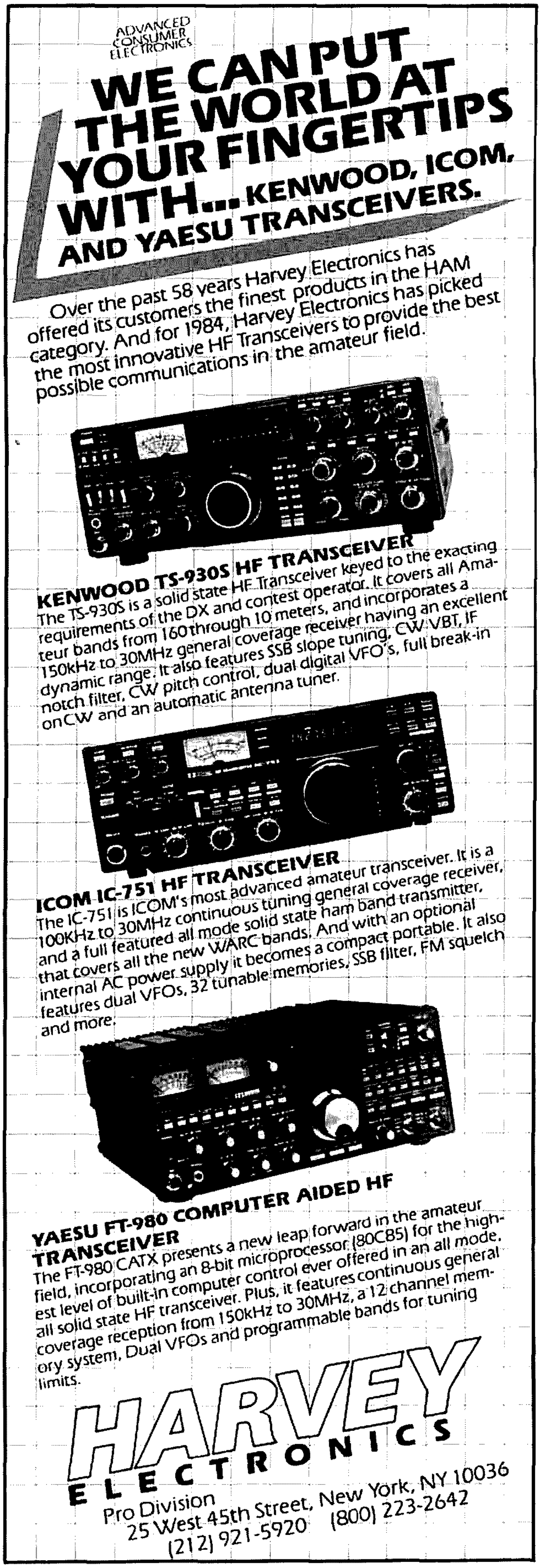


fig. 5. Guying details for various heights of Rohn 25 tower. (Reprinted with permission from Rohn Manufacturing, Peoria, Illinois.)



You will also need the initial distance from the neutral axis to the extreme fiber where failure occurs, c.

$$c = \frac{d_l}{2} \text{ inches} \tag{5}$$

The flexual strength can then be obtained by

$$f = \frac{12\ M\ c}{I} \text{ lbs./in.}^2 \text{ (psi)} \tag{6}$$

The constant 12 is necessary for the feet-to-inch conversion if the moment is expressed in foot pounds.

When the flexural strength exceeds the yield strength given for the particular member, the member will deform or break. The yield strength for tubular steel is typically 50,000 - 100,000 pounds per square inch. You will need to check with your local supplier to obtain this information.

Your masts and supporting booms should be strong enough to support your antenna system and be able to withstand the environmental conditions in your area. For example, if you live near salt water or in a highly industralized area, you should be especially vigilant about preventing and correcting corrosion.

Normally antenna and tower manufacturers are aware of these concerns and take steps to protect their product. Antennas are made from a corrosion-resistant alloy of aluminum, usually 6063-T6 or 6061-T6. Towers are normally hot-dip galvanized steel.

Masts purchased locally may not have any protection at all. If the mast has a very thick wall, this may not be a problem, but thin-walled steel or aluminum may be susceptible to corrosion if not protected.

summary

It may be unwise to follow the old saying, "If it doesn't come down, it isn't big enough." It's to your advantage to ensure the integrity of your antenna/tower system by matching the components so that your antenna system stays where it's supposed to stay. Know the limitations of your system's components and how each component interacts with the other and with the environment. It's up to the manufacturer to supply you with sufficient information to enable you to analyze your system, so that you can enjoy your hobby and not have to worry about your installation or your neighbor's roof.

references

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3. Unarco/Rohn Tower Catalog, 1976 edition, Rohn, P O. Box 2000, Peoria, Illinois 61656.
4. EIA Standard: Structural Standards for Steel Antenna Towers and Antenna Supporting Structures, RS-222-C, Engineering Department, Electronic Industries Association, Washington, D.C., March, 1976.

ham radio

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TOWER INSTALLATION:

make it sturdy, make it safe

For beginner to expert — useful information for safe tower installation

It's no joke to wake up some morning after a bad storm, walk outside, and see all or part of your tower lying haphazardly on the ground. Even less amusing is the sight of your antenna lying in the yard next door. People can be hurt — even lose their lives — because a tower wasn't big enough for its job, or wasn't erected properly . . . or both. Careful consideration of several factors can help prevent disaster and enhance your peace of mind. These factors are the tower type and size, versus the load on top; the adequacy of the tower base; a proper guying system; safe guy anchors; and proper tower maintenance.

choosing the right tower

Your first and main concern is wind. It can do unbelievable things to a tower, and it's of prime importance that you have the greatest respect for that fact. A basic rule of thumb is that if there's the slightest doubt, make your tower *stronger* than the situation seems to merit. A little extra money spent on your tower may be the cheapest insurance you'll ever buy.

The force the wind exerts against your tower depends on the number of square feet of surface area presented to the wind by the tower and the antennas, masts, and rotors on top. The tower manufacturer provides a specification indicating the maximum load, in square feet of antennas and related equipment, that the tower can bear. This specification means, of course, that you don't have to add the square footage of the tower to the antenna load. In other words, it means that the tower is strong enough to handle the specified load as well as whatever square footage is presented by the tower itself.

In turn, the antenna manufacturer gives you the square footage presented to the wind by its antenna. What it all adds up to is that if the tower manufacturer says its tower will handle a 16-square foot load and the antenna specification is 12 square feet, you're on safe ground, even after adding the mast and rotator. If, however, antenna, mast, and rotator come to just 16 square feet, that might be a little too close for comfort.

Another important consideration when choosing the correct tower for maximum safety is where you live. The Electronic Industries Association has done quite a bit of research, assembling meteorological data, county by county, throughout the United States, specifying the maximum wind likely to be encountered in any given location.

The United States (as well as most other areas on earth) is divided into three zones: Zone A, with winds up to 87 MPH (140 km/hr); Zone B, with winds up to 100 MPH (161 km/hr); and Zone C, with winds up to 112 MPH (180 km/hr). If the tower has been manufactured to meet standards established by the Electronic Industries Association, the manufacturer will specify, along with the square footage of the antenna's wind load, in which zone that square footage will apply. Of course, this specification assumes that the tower owner has followed the tower manufacturer's rules for installation and has installed an appropriate base and guy system.

It's a good idea to question the supplier and/or manufacturer before making your final selection. If you choose the correct tower and install it according to the manufacturer's recommended procedure, any reputable manufacturer should stand behind its product. If you want to learn about tower selection, installation, and maintenance, contact either a dealer or a tower manufacturer and obtain a copy of a complete catalog. Rohn, for example, provides a catalog (for a modest charge) that contains a wealth of valuable information applicable not only to Rohn towers but to towers in general.*

the tower base

It's important to determine whether the base will be used for a guyed tower or a self-supporting tower — and there's a big difference. In the case of the guyed tower, the force of the wind against the tower is translated by the guys from a lateral force into a

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downward compression. This places that force plus the dead weight of the tower *on the base*, with the initial tension of the guys also included in the total "weight" to be supported by the base.

The self-supporting tower poses a totally different problem, as far as the base requirement is concerned. Without guys to offer added support, the concrete base on which the tower stands must be much larger, just to keep the wind from uprooting the tower as it might uproot a tree and its entire root structure.

In the absence of guys, the wind force against the tower exerts downward pressure *on the legs* of the tower opposite the side from which the wind is blowing. Equally important is a strong upward lifting force on the legs nearest the source of the wind. *This means that the manner in which the tower is anchored into the base is very important*. A frequent cause of tower failure comes from trying to use a tower designed as a guyed tower as a self-supporting tower. Somehow, the reasoning seems to be that since Rohn 45G sections, for example, are quite strong, it should be OK to put up 40 or 50 feet without guys and mount a triband beam up there. But the base arrangement was simply not designed to bear this kind of load. Oh, I know, somebody out there is saying, "What's he talking about? That's exactly what I'm doing." My answer is, "I hope you're supporting only a 2-meter antenna, that you live in Zone A, and that the fates are kind!" I've seen people get away with — for a while — tower setups that would give a good mechanical engineer nightmares. Always review your tower plan with an experienced tower builder or mechanical engineer. It probably won't cost much — if anything at all, and may save a lot.

Follow the manufacturer's recommendations for your base. Use steel reinforcement bars ("rebar") to strengthen the base. Keeping the steel bars sealed inside the concrete will minimize rust problems. *It is extremely important to keep all ground rods outside and at least 4 inches from the base, bonding these rods directly to the steel tower itself*. If a ground rod goes through the concrete base, lightning can split the base in its search for a path to ground.

guy systems

Most guy systems consist of three guys at each of several levels, equally spaced around the tower and equal in length, with sets of guys spaced up the tower at intervals no greater than 30 feet (even closer if the tower is heavily loaded). For a three-guy system, the guys should ideally be anchored at a distance equal to 80 percent of the tower height. For a light load, such as a small tribander, this distance can be reduced, to say, 65 percent of the tower height. Where you cannot extend the guys this far, or where there is less available space on one side of the tower, you can reduce the distance from the tower to the anchor to as little as 40 percent of the tower height by using four guys at each level. However, as with three guys, their lengths and the spacing between them must be perfectly symmetrical.

Incorrect choice of guy wire is a frequent contributor to tower failure. Somehow the rumor has gotten around that aircraft control cable is suitable for use as guy wire. While it will rust, is harder to work with, and is about 25 percent weaker than real guy wire, you can get by with it . . . but do you really want to? The correct guy wire for most ham towers is 3/16 inch seven-strand EHS (extra high strength) wire. I recommend that you use nothing smaller than this. It will even handle some towers over 100 feet tall, depending on zone, and wind load. Follow the tower manufacturer's recommendation and you will find yourself using either 3/16 inch or for the really big ones, 1/4 inch EHS wire.

It's important, too, that these guys be tensioned properly. A good rule of thumb is to tension the wire to about 10 percent of its breaking strength. It should go without saying that the tension should be equal on all guys at a given level. The reason for using prestretched steel wire at the proper tension is that the tower must be prevented from twisting when the beam is buffeted by high winds. More towers are destroyed by twisting torque forces than in any other way; they just twist in the wind and collapse. For this reason, when you're using long-boom antennas or stacked monobanders, it's advisable to use special "guy assemblies" on the tower, providing stiff steel arms to improve the leverage the guys can exert in preventing the twist.

Be sure to use top-quality galvanized hardware for your guys. This is no place to save money, and the difference between the cheap kind and the rustproof kind is simply not worth the gamble. Use three clamps at each junction, and put the "U" of each clamp over the short, or so-called "dead-end" side of the cable. Use "thimbles" (those horseshoe-shaped pieces) which prevent the cable from kinking when it goes through the eye of a turnbuckle or a hole in a steel plate. Be sure to use the correct turnbuckle for the size of the guy you're using. Don't scrimp on turnbuckles; be sure they're rustproof and that the eyes are forged, not cast. Finally, when you've tightened up the turnbuckles, run a piece of guy wire through the body and both eyes of the turnbuckle in a "figure-eight" configuration to prevent the turnbuckle from working loose. If more than one guy terminates at such a point, use only one figure-eight wire, passing it through all the eyes and turnbuckle bodies.

Connecting the other ends of these guys to the tower is generally accomplished in either of two ways. When using the guy assemblies previously discussed, the guys are looped through the holes in the ends of

*UNR Rohn, Inc., Box 2000, Peoria, Illinois 61656.

the torque arms, using thimbles to prevent kinking; if there are no guy assemblies and the tower members are tubular (such as Rohn), take a single turn around the vertical member, where the cross-members intersect the vertical part. Be sure the loop also goes around the bends in the cross-members, encompassing both bends and the vertical part, compressing them instead of pulling them apart. After making the single turn around the tower in this fashion, fasten with three clamps as usual.

the anchors

Conventional anchoring for most Amateur towers employs the screw-type anchor. This is a 4-foot by 5/8-inch steel rod with a forged eye at one end and a 6-inch diameter auger on the other end. This is simply screwed into the ground at the same angle taken by the guy wire that will be attached to it. This makes a surprisingly strong anchor for some pretty sizable towers. For example, the 68-foot Rohn 45G fold-over tower uses this one. In any case, consult the manufacturer's recommendation. The other anchor normally used when the screw-type is not strong enough is the concrete anchor. This is a 5-foot by 5/8-inch steel rod, with a forged eye on one end and a crook on the other end, encased in concrete. When two or more guys terminate at the same anchor, equalizing plates are often used to divide the tension equally between the guys. Through-the-wall anchors and other types are available for those cases where restricted space does not provide room for conventional ground anchors. Again, the big Rohn catalog describes every conceivable kind of special hardware and includes detailed instructions on how to use such items.

tower maintenance

Most of this is just common sense and we really know pretty much what to do. But time flies and we never seem to get around to maintenance as often as we should. We should check our guys for equal and proper tension. When checking for equal tension, we should use a long carpenter's level, placing it against the vertical part of the tower to be sure that it is truly vertical, or "plumb." A perfectly vertical tower is not affected by the unusual stresses invariably present when a tower is off "plumb." All bolts and nuts should be checked for tightness and replaced if they are rusty. Anything that shows rust but cannot be replaced (such as the tower itself), should be treated. This can be done most effectively with a cold-galvanize spray, which will form a chemical bond with the good galvanizing around the rust spot and prevent further rust formation, making the spot virtually as good as new. There are several brands of this material; the one I have used with good success is made by LPS.*

*LPS® Cold Galvanize, Holt/Lloyd LPS, 4647 Hugh Howell Road, Tucker, Georgia 30084.

conclusion

Be very careful about hinging towers at the base and "walking" them up. This may appear to be the easiest way to erect 40 or 50 feet of tower, but more often than not, it is the most difficult and most dangerous. If you have plenty of people present — at least one of whom understands mechanical stresses — proceed carefully. If not, you may find it easier for just two people to put that tower up, one on the tower with a safety-belt and a "gin-pole," and the other on the ground to handle the pulley rope and pick up the tools the other drops.

In regard to the mast you put in the top of the tower to support your beam, be aware that if the height of the mast above the tower is, say, 10 feet, and the wind is blowing only 60 miles an hour against 8 square feet of antenna, there are some horrendous forces trying to convert that mast to rubble. *You'll be safest with a good steel mast.* A good aluminum-alloy mast may be lighter and just as strong right up to the instant that it crystalizes — something that doesn't happen to a tough steel mast.

High-quality close-woven nylon can do a good job when used for the right job — for instance, for supporting masts or the ends of dipoles. But for towers, it's a different story: twisting destroys towers so, very simply, nylon guys will stretch enough to let them twist; steel guys won't. Incidentally, when guying masts, use only close-woven nylon. The older it gets, the tougher it gets, but the hot sun will eventually disintegrate Dacron and other synthetics.

One more comment on anchors and bases: all that I've said so far has been predicated on the assumption that you'll be constructing your tower over normal soil. When the soil is swampy or sandy, consult an experienced tower builder or a civil engineer.

A couple of notes on climbing: *Spend the money for a good quality, approved safety belt.* It's cheap insurance. (A friend and former colleague lost his life using a home-made safety belt.) When climbing a tower, you come to the moment of truth every time you have to unhook the belt to move up around a set of guys. It may be a little more trouble but it's a good idea to equip your belt with *two* lanyards. You can then leave one fastened around the tower below the guys until you've fastened the other one above the guys . . . after which you can unfasten the lower one and move up with complete safety. If you're squeamish about climbing towers, this system will provide both safety and reassurance.

An excellent source of information on this subject is "Structural Standards for Steel Antenna Towers and Antenna Supporting Structures," *EIA Standard RS-222-C*, March, 1976, available from EIA, 2001 Eye Street, N.W., Washington, D. C. 20006, at $7.40.

ham radio

TOWER MAINTENANCE:

keep your tower *UP*

Regular maintenance means lasting success

Those of us fortunate enough to have towers with moderate to complex antenna systems sometimes take them for granted. Even though we've invested substantial amounts of time and money in planning and erecting our systems, once they're up we tend to head for the ham shack and forget that they're there. Only the rotator control and the S-meter remind us that there's something out there, after all.

That's why I've designated a specific time of year — well before winter sets in — for a thorough annual inspection of every inch of metal in my system.

Good planning keeps the number of trips up the tower to a minimum, and the work pleasurable.

Choose a comfortable time of year; windy, hot or cold days will only discourage you from staying on the tower any length of time. I usually do my inspection in the fall, when temperatures are moderate and it's a joy to be up there. There's a built-in benefit for your system, too, in choosing fall as the time for your annual inspection. After a summer of heat and ultraviolet stress — and battering by the winds of thunderstorms and perhaps even hurricanes, the stiffening temperatures of winter can bring tape and cables to their natural end. Those who live in areas where seasonal changes are less severe may opt for a different time.

inspect the antenna

Gather the appropriate tools (pliers, wrenches, tape, sealant, etc.) and an approved (not home-made) safety belt. Climb up once, check and correct any deficiencies on the way *down*, and celebrate your good planning.

Most of your time will be spent right at the top, where you'll first make a visual inspection of the antennas. Although most of the hardware will be beyond your reach, a look at the general condition will reveal a great deal. Loose or missing hardware is a sure sign of trouble. Sometimes scratches and the general pattern of weathering will indicate any elements turned from original position. After your visual inspection is complete, *shake* the whole assembly; you'll hear or see anything that's come loose.

The first part of the beams to deteriorate is often the support cable for the boom. Check that thoroughly, because without it the wind tolerance of your beam may be far less than you think. All broken or missing parts should be replaced, even if that calls for a major antenna party. Electrical connections should also be checked thoroughly, since an increase in the resistance at the feedpoint can mean needless loss of power. I routinely spray all connections with clear acrylic sealer. Available in the spray paint section of most hardware stores, this product will prevent corrosion on the connections. A saturating coat sprayed over all connections once every two or three years is easy to apply and seals all cracks, yet allows disassembly when necessary.

don't forget the feedline

While at the top of the tower, remember to check the attachment of the feedline to the antenna and mast. This provides strain relief for the antenna connection. Tight taping is normally used here, but even the stresses of normal turning can loosen these support points. Although most of us use electrical tape for this, fiberglass reinforced strapping tape will serve

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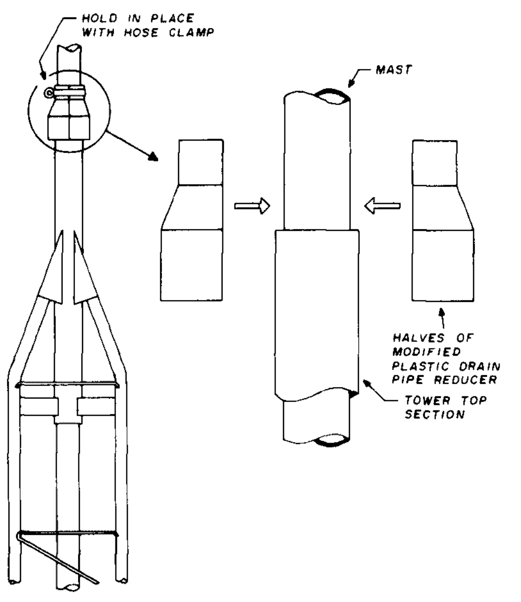

fig. 1. Illustration of split drain pipe reducer as drip-cap for tower top.

well at high-stress points such as this. If loose, retaping is in order; pay particular attention to the loop of feedline necessary to allow rotation of the antenna. Multiple feedlines can be taped together to support each other, but a larger loop should be allowed to accommodate the additional stiffness. (If you put the antenna at the center of its rotation *before* climbing the tower, it will be easier to visualize where this rotation loop must be located, and any necessary repair will thus be easier to spot and simpler to do.)

grease fittings: good idea

A step down the tower's top section will allow inspection of the mast bearing. Check and replenish the lubrication of this bearing if necessary. Most of us use a top section that has no true bearing, utilizing a top pipe section to guide the mast instead; these sections are particularly difficult to lubricate, since the pipe length may be as long as 30 inches. The mast in most installations like this fits loosely, with approximately 0.1 inch (2.5 mm) diametrical clearance. Using a hand grease gun to inject grease through the various holes already in the tower pipe makes this chore easier, and a homemade rubber washer will serve as a seal when pressed over the hole with the grease gun while lubricating.

To simplify lubrication of the entire length of the pipe, I installed automotive grease fittings in a few places along the tower pipe. If you do this, take care to choose fittings that are flat on their bottoms, and install them so they don't protrude into the inside of the tower pipe, because this would cause a wear point on the mast. One alternative is to leave the fittings in only during the lubrication process and then cover the holes with tape. I admit that while installation of the fittings is a cinch on a new tower still on the ground, it could be quite a challenge on an existing tower.

Water and dirt must be kept out of the bearing area. If they're not, your antenna may freeze in place during the winter. You can easily install a drip cap to prevent this, using readily available hardware.

My mast pipe, like most, is a standard pipe size. For the typical installation, 1.5 inch (38.1 mm) pipe which is 1.9 inch (48.3 mm) OD fits through a 2 inch (50.8 mm) ID tower tube, which is in turn about 2.25 inches (57.2 mm) OD. Since the mast pipe is a standard pipe size, a 1.5 inch (38.1 mm) to 2 inch (50.8 mm) reducer for plastic drainpipe makes an excellent drip cap (see **fig. 1**). Just cut it in half with a hacksaw, file out the internal chamfered stop, and fit the two halves snugly over the mast pipe. Use an automotive hose clamp to hold it in place over the top of the tower pipe; water and dirt will be excluded, and the mast will rotate freely for a longer length of time than it might without protection (see **fig. 1**). For other standard mast pipe sizes, different size reducers are required; there should be no problem as long as the water pipe used for the mast and the plastic drain pipe follow the same size conventions.

on the way down

On your way down, check all cables for proper support, replacing tape as necessary. Look for any rust; it's a sure sign that galvanizing in that spot on the tower has been scraped off. Remove the rust with a light sanding and seal with acrylic sealer or aluminum paint.

Check bolts and nuts for tightness, both at the tower section joints and on any other hardware. Guy wire attachment points are particularly important. Treat all bolts to prevent their nuts from rotating off should loosening occur; this can be done by striking a center punch against the bolt thread protruding beyond the nut. By slightly upsetting the thread, you'll prevent the nut from vibrating off, but don't be so aggressive that you destroy both the thread and your chances of disassembly later on.

on the ground

Back on ground level, check the cable entry into the house to make sure it is still properly sealed. Check the connections to the tower ground system, too, both for tightness and for any signs of corrosion that would cause poor contact. If the grounding wires have become kinked for any reason, straighten them as needed for optimum protection against lightning.

Now take a walk around the yard. Are all guy wires and turnbuckles secure, all nuts and bolts tight? Here in particular, upsetting the threads of the bolts as described above is appropriate, since they generally can't

be tightened completely and still allow equalizer plate motion. Note the condition and tension of the guy wires: any signs of rust or wear in the wire calls for immediate replacement. In checking tension, remember that it will change from season to season with rising and falling temperatures, which is why I check mine in the fall when temperatures are moderate. This factor is important enough for tower manufacturers to recommend different guy wire tension for different temperatures at the time of installation. Lacking the equipment needed to measure tension, a rule of thumb is to tighten the turnbuckles as much as you can with your bare hands. This will normally leave some sag in the wires, especially if insulators are installed, but don't worry. The last bit of sag requires a great deal of tension to remove and puts undue stress on the wires and the tower. Remember that the wires are pulling not only *out* but also *down*, and all that force has to be supported by the tower.

Finally, if you haven't installed safety loops and checked them, install one now at each guy anchor. A safety loop may be as simple as a short loop of guy wire threaded through the guy wire ends and the anchor rod loop, with the ends held together with normal wire clamps. The loop serves to catch the guy wires and save the tower if a turnbuckle should break. The loop may also be threaded through the turnbuckles in figure-eight fashion to keep the turnbuckles from loosening.

ham radio

applied Yagi antenna design
part 2:
220 MHz and the Greenblum design data

220 MHz Yagis with different tapers provide best gain and F/B figures

This article addresses antenna design for the 220 MHz VHF band. In comparison to 144 MHz, this band offers similar propagation conditions, with a reduction in antenna size of approximately 35 percent. Hence, for the same physical boom length, more gain and a sharper pattern can be realized. The antennas described for 220 MHz differ from those analyzed for 144 MHz[1] for another reason as well: the Greenblum antenna design data, rather than the Kmosko-Johnson data used in last month's article, is used for the 220 MHz Yagis.

If widespread and long term publishing and republishing are reliable indicators of popularity, then Greenblum's set of charts and curves must be one of the best known design tools in all of Amateur Radio. This formerly Telrex-proprietary data was first published in 1956 as a two-part series.[2,3] One of the first non-Telrex VHF applications was Tilton's six-element 50 MHz Yagi.[4] Applications for other VHF and UHF bands followed.[5] A recent computer-optimized 144 MHz moonbounce antenna developed by Joe Reisert, W1JR, was also based on Greenblum's data.[6]

Yet, as is the case for the 220 MHz band to which this article is addressed, Greenblum's design data remains somewhat of an enigma. Neither Greenblum nor those who republished his charts and tables ever specified a reflector length. Reflector spacing is given, as are director lengths and spacings, but no reflector lengths are provided. As part of the preliminary work in preparing this article, Greenblum antennas using mid-range element spacing values were modeled. With a 100:1 wavelength-to-diameter ratio, maximum gain with six elements occurred with a reflector length of 0.49 wavelengths. The overall pattern was, however, of little use. Reflector lengths dropped as element numbers were reduced, until a reflector of 0.479 wavelengths was reached for the two-element beam. As these antennas would be of little use at 220 MHz, their modeling was not pursued any further.

technical background

As was the case with Lawson,[7] Greenblum's design calculations were derived from Uda and Mushiake.[8] Greenblum's reactance charts and his formula are referenced to the very same pages in Uda and Mushiake's work as those to which Lawson referred. In 1978, Powers[9] reconfirmed the basic Greenblum data by using element spacings from the middle of the specified ranges for all parasitics. Gain and F/B figures similar to those originally stated by Greenblum were said to have been obtained.

Tilton's previously cited work produced a different set of element lengths and spacings. He found that the more or less traditional tapering schemes in use for VHF/UHF antennas resulted in an increase in forward gain. Element spacings used were towards the lower end of the specified range. When the Tilton/Greenblum designs are compared against Viezbicke's findings,[10] very favorable boom-to-gain ratio comparisons are realized. One further measure of the value of Tilton's modifications was the use of his 50 MHz design by a well known commercial antenna manufac-

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table 1. Reference Tilton/Greenblum 220.5 MHz antenna with specified element spacings and parasitic element lengths to be supplied with each computer iteration.

	element	length (inches)	cumulative spacing (inches)	cumulative spacing (λ)
1	Reflector	—	0.0	0.0
2	driven element	25.686378	7.5	0.14011
3	director 1	—	15.0	0.28022
4	director 2	—	24.0	0.44836
5	director 3	—	36.5	0.68188
6	director 4	—	50.5	0.94343
7	director 5	—	65.0	1.21431
8	director 6	—	83.0	1.55058
9	director 7	—	101.0	1.88685
10	director 8	—	119.0	2.22312
11	director 9	—	137.0	2.55939
12	director 10	—	155.0	2.89566
13	director 11	—	173.0	3.23193

turer. In view of Tilton's reputation and the widespread availability of his lifelong work in VHF/UHF antenna applications, a Tilton/Greenblum design will serve as the basis for computer iteration of a 220 MHz Yagi antenna.

the Tilton/Greenblum design

As a result of his previously cited work with the Greenblum data, Tilton published documentation on two 220 MHz antennas. An eleven-element Yagi appeared in the 1968 *Radio Amateur's Handbook*,[11] and a seven-element Yagi appeared in *The Radio Amateur's VHF Manual*.[12] The longer design is a scaling and reoptimization of a 432 MHz antenna of eleven elements. Based on the nature of the Greenblum data, these would appear to be two very different 220 MHz antennas.

For purposes of computer iteration, the eleven-element antenna is extended to thirteen elements by adding two more directors. **Table 1** shows the reference design upon which this article is based. Since it is predicated on the Greenblum design data, the tapering scheme continues through the additional directors, and their spacing is identical to the spacing of the ninth director.

The design frequency is 220.5 MHz, which allows optimizing the design to cover the weak signal area of the 220 MHz band. Yagis at this frequency are more inherently broadbanded than at 144 MHz. With few exceptions, the designs optimized at 220.5 MHz will work equally well over the first few megahertz of the entire band. There are strong indications that Tilton's eleven-element design was optimized close to this design frequency, and that the seven-element Yagi was optimized at 221.5 MHz.

table 2. Optimized gain iteration for a 0.000 taper with a 26.375-inch reflector.

director 1 (inches)	gain (dBi)	F/B(dB)
22.000	14.081	22.951
22.125	14.238	23.961
22.250	14.396	24.725
22.375	14.555	24.863
22.500	14.712	24.122
22.625	14.863	22.680
22.750	15.005	20.929
22.875	15.131	19.149
23.000	15.234	17.465
23.125	15.305	15.923
23.250	15.332	14.537
23.375	15.304	13.317
23.500	15.212	12.281
23.625	15.051	11.460
23.750	14.821	10.909
23.875	14.527	10.729
24.000	14.162	11.100

table 3. Optimized F/B iteration for a 0.000 taper with a 26.875-inch reflector.

director 1 (inches)	gain (dBi)	F/B(dB)
21.750	13.618	29.037
21.875	13.776	29.793
22.000	13.937	29.739
22.125	14.099	28.388
22.250	14.262	26.588
22.375	14.424	24.622
22.500	14.582	22.817
22.625	14.733	21.115
22.750	14.872	19.563
22.875	14.994	18.153
23.000	15.092	16.874
23.125	15.158	15.721
23.250	15.184	14.696
23.375	15.163	13.805
23.500	15.093	13.064
23.625	14.971	12.498
23.750	14.801	12.135
23.875	14.578	12.001
24.000	14.271	12.060

table 4. Frequency response parameters for the zero taper gain optimized antenna.

frequency	gain (dBi)	F/B (dB)
216.5	14.737	16.075
217.5	15.007	16.472
218.5	15.194	16.169
219.5	15.301	15.425
220.5	15.332	14.537
221.5	15.295	13.695
222.5	15.201	13.003
223.5	15.065	12.515
224.5	14.903	12.266

Tilton included matching techniques with the designs for both of his antennas. The iterated thirteen-element antenna can easily use the same matching hardware as the eleven-element antenna because the two additional directors are not going to change the antenna's input impedance by very much. And because Tilton used wooden booms for his antennas, element length conversion calculations are not necessary for the reference antenna. The element diameter remains at 0.125 inches. As 0.0625 inches represents the closest tolerance to which most Amateurs can cut antenna elements, tapering schemes will be expressed in multiples of 0.0625 inches. In terms of the parasitic element iteration lengths, multiples of 0.125 inches are used. No real differences were found at 220 MHz by using a finer increment.

computer-designed Tilton/Greenblum antennas

Tilton presented one long Greenblum antenna for 220 MHz. This antenna had a taper of 0.125 inches. For purposes of exploring this design via computer iteration, five antennas are presented. In increments of 0.0625 inches, their tapers vary from zero to 0.25 inches. Each antenna is presented with tables showing optimized gain and F/B iterations, calculated performance across an 8-MHz bandwidth (216.5 MHz -

table 5. Frequency response parameters for the zero taper F/B optimized antenna.

frequency	gain (dBi)	F/B (dB)
216.5	13.436	19.360
217.5	13.533	21.773
218.5	13.620	24.604
219.5	13.700	27.743
220.5	13.776	29.793
221.5	13.852	28.567
222.5	13.928	25.867
223.5	14.004	23.402
224.5	14.079	21.381

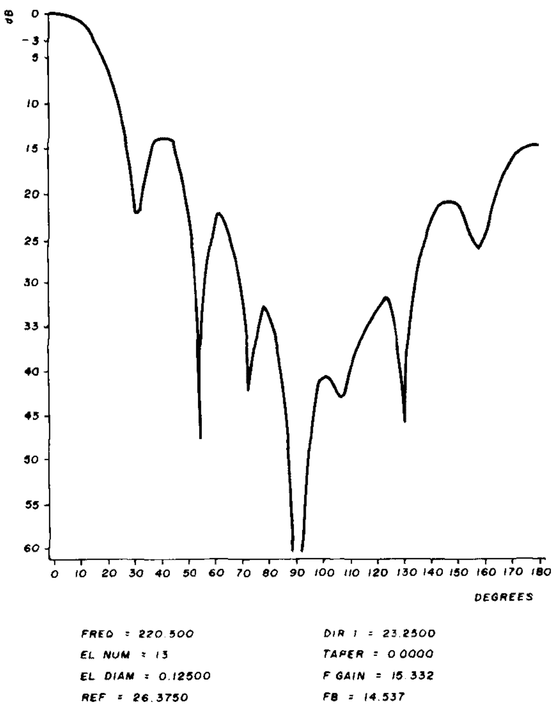

fig. 1. Gain optimized antenna with a zero taper.

table 6. Optimized gain iteration for a taper of 0.0625 with a 26.375 inch reflector.

director 1 (inches)	gain (dBi)	F/B (dB)
22.000	13.745	20.212
22.125	13.894	20.984
22.250	14.046	21.837
22.375	14.201	22.748
22.500	14.358	23.639
22.625	14.515	24.341
22.750	14.671	24.582
22.875	14.824	24.141
23.000	14.971	23.063
23.125	15.108	21.617
23.250	15.230	20.063
23.375	15.331	18.553
23.500	15.402	17.161
23.625	15.437	15.924
23.750	15.427	14.868
23.875	15.366	14.025
24.000	15.250	13.446

table 7. Optimized F/B iteration for a taper of 0.0625 with a 26.875 inch reflector.

director 1 (inches)	gain (dBi)	F/B (dB)
22.000	13.597	27.636
22.150	13.753	28.716
22.250	13.912	29.485
22.375	14.073	29.543
22.500	14.235	28.702
22.625	14.397	27.227
22.750	14.558	25.514
22.875	14.714	23.809
23.000	14.862	22.211
23.125	15.000	20.753
23.250	15.121	19.437
23.375	15.221	18.266
23.500	15.294	17.241
23.625	15.334	16.372
23.750	15.337	15.679
23.875	15.301	15.194
24.000	15.224	14.969

table 8. Frequency response parameters for the 0.0625 taper gain optimized antenna.

frequency	gain (dBi)	F/B (dB)
216.5	14.748	16.050
217.5	15.096	16.967
218.5	15.280	16.992
219.5	15.391	16.552
220.5	15.437	15.924
221.5	15.426	15.328
222.5	15.372	14.904
223.5	15.287	14.750
224.5	15.180	14.954

fig. 2. F/B optimized antenna with a zero taper.

224.5 MHz) for each optimized antenna, and cartesian plots of each antenna.

taper = 0.000

Table 2 presents the gain optimizing iteration that resulted in 15.332 dBi of gain, and **table 3** presents the F/B optimizing iteration and its calculated result of 29.793 dB of F/B. Differences of over 1.4 dB in gain and 15 dB in F/B ratio exist between these antennas. **Tables 4** and **5** present these antennas' respective calculated performance over the specified bandwidth. Both antennas show marked peaks at 220.5 MHz in their respectively optimized parameters. **Figs. 1** and **2** present these antennas' respective E-plane plots. The differences in main lobe width and depth are readily apparent, as is the difference in signal attenuation from 160 to 180 degrees. It is interesting to note the obvious differences between optimized antennas with a zero taper when both antennas are based on a design approach that requires a measureable taper.

table 9. Frequency response parameters for the 0.0625 taper F/B optimized antenna.

frequency	gain (dBi)	F/B (dB)
216.5	13.694	20.134
217.5	13.802	22.671
218.5	13.898	25.607
219.5	13.988	28.547
220.5	14.073	29.543
221.5	14.156	27.645
222.5	14.238	25.084
223.5	14.318	22.869
224.5	14.396	21.061

table 10. Optimized gain iteration for a taper of 0.125 with a 26.875 inch reflector.

director 1 (inches)	gain (dBi)	F/B (dB)
22.000	13.286	24.765
22.125	13.434	25.662
22.250	13.586	26.654
22.375	13.740	27.707
22.500	13.898	28.715
22.625	14.057	29.450
22.750	13.325	18.066
22.875	14.380	28.934
23.000	14.540	27.676
23.125	14.698	26.146
23.250	14.850	24.586
23.375	14.994	23.113
23.500	15.127	21.774
23.625	15.243	20.587
23.750	15.339	19.563
23.875	15.410	18.718
24.000	15.543	18.078
24.125	15.556	20.338
24.250	15.439	17.632
24.375	15.371	18.042
24.500	15.243	19.164
24.625	15.014	21.422
24.750	14.608	24.627
24.875	13.952	22.689
25.000	13.093	17.932

taper = 0.0625

Table 6 presents the gain optimizing iteration that resulted in 15.437 dBi of gain, and **table 7** presents the F/B optimizing iteration and its calculated result of 29.543 dB. Nearly 1.4 dB of gain and over 13.6 dB of F/B separate these antennas. **Tables 8** and **9** present these antennas' calculated performance over the specified bandwidth. Both antennas show peaks at

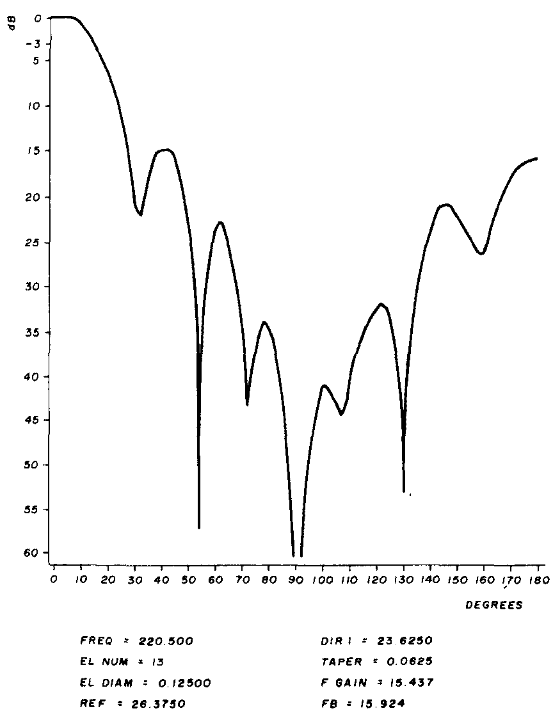

fig. 3. Gain optimized antenna with a 0.0625 taper.

fig. 4. F/B optimized antenna with a 0.0625 taper.

table 11. Optimized F/B iteration for a taper of 0.125 with a 26.375 inch reflector.

director 1 (inches)	gain (dBi)	F/B (dB)
22.000	13.441	18.614
22.125	13.581	19.146
22.250	13.725	19.740
22.375	13.873	20.402
22.500	14.022	21.136
22.625	14.175	21.935
22.750	14.329	22.773
22.875	14.483	23.580
23.000	14.638	24.213
23.125	14.790	24.465
23.250	14.938	24.161
23.375	15.079	23.310
23.500	15.208	22.106
23.625	15.322	20.768
23.750	15.415	19.450
23.875	15.480	18.242
24.000	15.511	17.198
24.125	15.504	16.363
24.250	15.452	15.793
24.375	15.352	15.575
24.500	15.195	15.862
24.625	14.957	16.951
24.750	14.597	19.543
24.875	14.052	26.141
25.000	13.318	32.565
25.125	12.599	21.157

table 12. Frequency response parameters for the 0.125 taper gain optimized antenna.

frequency	gain (dBi)	F/B (dB)
216.5	15.371	20.118
217.5	15.444	19.343
218.5	15.479	18.549
219.5	15.484	17.955
220.5	15.556	20.338
221.5	15.423	17.879
222.5	15.396	18.630
223.5	15.257	20.619
224.5	15.067	24.350

220.5 MHz for their respectively optimized parameters. **Figs. 3** and **4** present these antennas' respective E-plane plots. As was the case for the zero taper antennas, there are obvious differences in main lobe width and depth, and in signal attenuation over the rear-most 20 degrees. The slight taper does not seem to have made much difference in the comparisons between the two 0.0625 antennas and the zero taper antennas. Both gain optimized antennas have the same reflector length, and this is also true of the two F/B optimized antennas. The 0.0625 antennas both have longer director lengths than their zero taper counterparts. Because element spacing has remained constant, this difference is due to director tapering.

fig. 5. Gain optimized antenna with a 0.125 taper.

fig. 6. F/B optimized antenna with a 0.125 taper.

table 13. Frequency response parameters for the 0.125 taper F/B optimized antenna.

frequency	gain (dBi)	F/B (dB)
216.5	14.650	12.560
217.5	14.489	14.049
218.5	14.264	16.806
219.5	13.903	22.937
220.5	13.318	32.565
221.5	12.630	20.061
222.5	12.274	17.038
223.5	12.411	14.964
224.5	12.412	13.429

table 14. Optimized gain and F/B iteration for a taper of 0.1875 with a 27.0 inch reflector.

director 1 (inches)	gain (dBi)	F/B (dB)
22.875	14.004	29.444
23.000	14.165	29.547
23.125	14.326	29.052
23.250	14.487	28.061
23.375	14.646	26.802
23.500	14.800	25.472
23.625	14.947	24.189
23.750	15.084	23.012
23.875	15.209	21.973
24.000	15.317	21.096
24.125	15.404	20.405
24.250	15.467	19.941
24.375	15.501	19.768
24.500	15.598	21.030
24.625	15.455	20.831
24.750	15.341	22.712
24.875	15.121	26.857
25.000	14.743	37.306
25.125	14.191	26.858
25.250	13.580	21.249

taper = 0.125

Table 10 presents the gain optimizing iteration that resulted in 15.556 dBi of gain, and **table 11** presents the F/B optimizing iteration that resulted in 32.565 dB of F/B. Over 1.2 dB of gain and over 12 dB of F/B separate these two antennas. **Tables 12** and **13** present these antennas' respective calculated performance over the specified bandwidth. Both antennas show easily determined peaks at 220.5 MHz in their respectively optimized parameters. The high F/B figure is the result of significant single frequency vectorial cancellation. A very good F/B will be recognized over the entire weak signal band segment. **Figs. 5** and **6** present these antennas' respective E-plane plots. The differences in main lobe width and depth are major. In comparison, the F/B optimized antenna almost does without a clearly defined main lobe, and its increased signal attenuation from 170 to 180 degrees comes at a nearly 10 dB (average) reduction in signal attenuation from 120 to 165 degrees. The high degree

table 15. Frequency response parameters for the 0.1875 taper gain optimized antenna.

frequency	gain (dBi)	F/B (dB)
216.5	15.449	21.271
217.5	15.503	20.549
218.5	15.528	19.973
219.5	15.527	19.734
220.5	15.598	21.030
221.5	15.442	21.004
222.5	15.330	23.280
223.5	15.124	28.150
224.5	14.765	30.512

table 16. Frequency response parameters for the 0.1875 taper F/B optimized antenna.

frequency	gain (dBi)	F/B (dB)
216.5	15.520	18.340
217.5	15.547	19.245
218.5	15.337	21.196
219.5	15.118	25.498
220.5	14.743	37.306
221.5	14.184	25.206
222.5	13.543	19.359
223.5	13.117	17.571
224.5	13.179	18.989

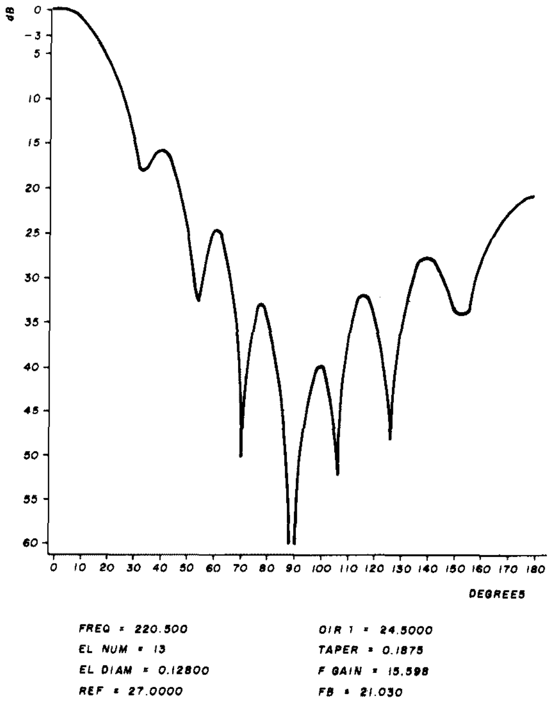

fig. 7. Gain optimized antenna with a 0.1875 taper.

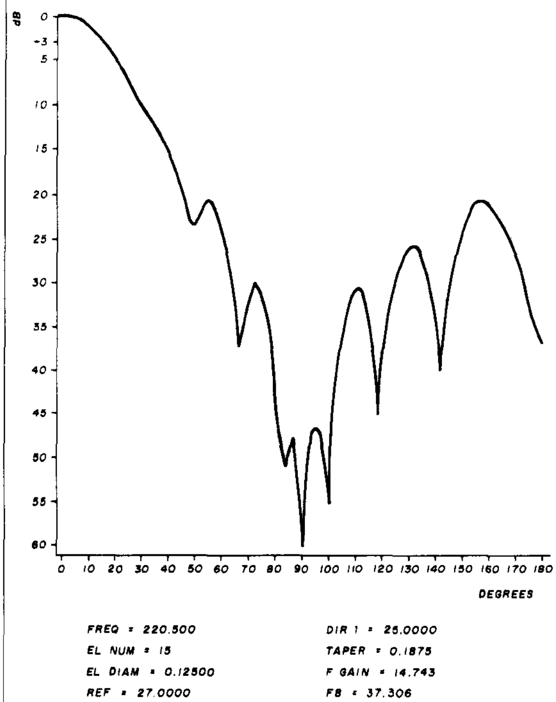

fig. 8. F/B optimized antenna with a 0.1875 taper.

of optimization of a single parameter comes at a comparative cost in performance over the rest of this antenna's pattern. Tilton's selection of a gain optimized antenna of this taper is soundly based on his own actual measurements and what this model's calculations have again realized. The F/B level realized by the gain optimized antenna produces a sharp pattern as well as an F/B ratio easily in keeping with the 220 MHz band's level of activity. In comparison with the zero taper antenna and the 0.0625 taper antennas, the 0.125 antenna has a longer first director. As element spacing is fixed, this difference is due to the increased director tapering.

taper = 0.1875

Table 14 presents the gain optimizing iteration that resulted in 15.598 dBi of gain, and it is also the F/B optimizing iteration that resulted in 37.306 dB of F/B. Just over 0.85 dB of gain and just under 16.3 dB of F/B separate these two antennas. **Tables 15** and **16** present these antennas' calculated performance over the specified bandwidth. Both antennas show easily located peaks at 220.5 MHz in their respectively optimized parameters. As was the case for the 0.125 taper antenna, this F/B optimized antenna's high F/B figure is the result of single frequency vectorial cancellation.

table 17. Optimized gain iteration for a taper of 0.25 with a 26.625 inch reflector.

director 1 (inches)	gain (dBi)	F/B (dB)
23.000	13.976	24.234
23.125	14.128	25.153
23.250	14.282	26.147
23.375	14.435	27.156
23.500	14.589	28.037
23.625	14.740	28.539
23.750	14.887	28.406
23.875	15.028	27.614
24.000	15.160	26.403
24.125	15.279	25.055
24.250	15.382	23.755
24.375	15.464	22.599
24.500	15.519	21.638
24.625	15.543	20.910
24.750	15.529	20.458
24.875	15.467	20.331
25.000	15.342	20.580
25.125	15.133	21.206
25.250	14.823	22.051
25.375	14.441	22.903
25.500	14.114	24.715
25.625	14.030	28.976
25.750	14.062	19.353
25.875	13.857	16.152
26.000	6.991	4.234

table 18. Optimized F/B iteration for a taper of 0.25 with a 26.875 inch reflector.

director 1 (inches)	gain (dBi)	F/B (dB)
23.000	13.901	27.711
23.125	14.057	28.746
23.250	14.214	29.606
23.375	14.372	30.219
23.500	14.530	30.137
23.625	14.685	29.369
23.750	14.835	28.147
23.875	14.980	26.760
24.000	15.115	25.397
24.125	15.237	24.153
24.250	15.344	23.072
24.375	15.429	22.182
24.500	15.490	21.509
24.625	15.519	21.093
24.750	15.510	20.990
24.875	15.450	21.272
25.000	15.320	22.011
25.125	15.095	23.134
25.250	14.759	24.066
25.375	14.351	24.140
25.500	14.030	24.943
25.625	13.994	27.748
25.750	14.026	18.951
25.875	13.842	16.766
26.000	5.770	3.280

Here too, a very fine F/B will be realized over the entire weak signal area. **Figs. 7** and **8** present these antennas' respective E-plane plots. The 0.1875 antennas compare with one another in a manner similar to the 0.125 antennas. The high cost of the high F/B ratio is all too apparent. The gain optimized 0.1875 antenna has a clean pattern and a respectable F/B. Both 0.1875 antennas have the same reflector length but continue the trend toward longer director lengths with an increased taper. However, the 0.125 and 0.1875 F/B optimized antennas have initial directors of the same length. Their difference is the latter's longer reflector.

taper = 0.25

Table 17 presents the gain optimizing iteration that resulted in 15.543 dBi of gain, and **table 18** presents the F/B optimizing iteration that resulted in 30.129 dB of F/B. Slightly more than 1.1 dB of gain and nearly 10 dB of F/B separate these two antennas. **Tables 19** and **20** present these antennas' calculated performance over the specified bandwidth. Both antennas have easily located peaks at 220.5 MHz in their respectively optimized parameters. Unlike the two previous F/B optimized antennas, the 0.25 taper F/B optimized antenna does not have a high single frequency F/B, and maintains a near-optimized F/B across the entire weak signal area. **Figs. 9** and **10** present the 0.25 antennas' E-plane plots. Though the gain optimized antenna has the narrower main lobe, both antennas have clearly defined main lobes. This is in contrast to the pairs of antennas compared at the 0.1875 and 0.125 tapers. Along a similar vein, the great disparities noted in the signal attenuation characteristics between antennas of the two most recently presented tapers, exist only to a limited degree between the 0.25 taper antennas. While the gain optimized 0.25 taper antenna continues the trend to longer director lengths, the F/B optimized antenna significantly *reverses* this tendency.

summary

The computer iterations performed on a family of ten 220 MHz Tilton/Greenblum Yagis indicate that the user needs to have a clear understanding of his or her antenna requirements before making a selection. There are great differences between the gain and F/B optimized antennas within each tapering approach. Additionally, for each of the optimized antennas, the best value of the other *(non-optimized)* parameter generally occured at a frequency far removed from the design frequency of 220.5 MHz. Given the broadbanded nature of Yagis on this band, very little gain is lost during even extensive changes in frequency. For some of the F/B optimized antennas, there are marked penalties in *F/B* for even slight frequency changes.

table 19. Frequency response parameters for the 0.25 taper gain optimized antenna.

frequency	gain (dBi)	F/B (dB)
216.5	15.210	20.451
217.5	15.360	21.259
218.5	15.462	21.417
219.5	15.522	21.186
220.5	15.543	20.910
221.5	15.528	20.845
222.5	15.469	21.154
223.5	15.354	21.922
224.5	15.156	22.978

table 20. Frequency response parameters for the 0.25 taper F/B optimized antenna.

frequency	gain (dBi)	F/B (dB)
216.5	13.948	20.464
217.5	14.067	22.851
218.5	14.175	25.567
219.5	14.276	28.419
220.5	14.372	30.219
221.5	14.466	29.425
222.5	14.557	27.249
223.5	14.646	25.119
224.5	14.732	23.343

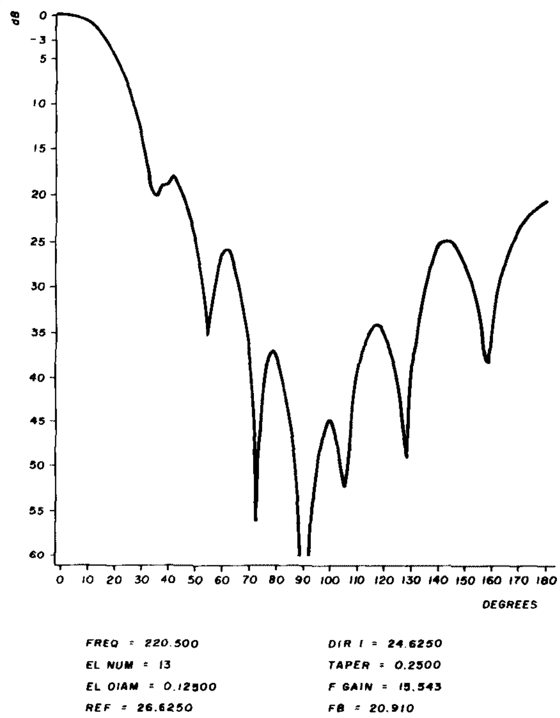

fig. 9. Gain optimized antenna with a 0.25 taper.

With the exception of the zero, 0.0625, and 0.25 taper antennas, F/B optimization is very clearly the result of single frequency vectorial cancellation. A user interested in reasonably high F/Bs that will be recognized across the entire weak signal area (of the band) could easily choose from among the various 0.0625 and 0.25 taper F/B optimized antennas. There is the added bonus of reasonably good gain figures and clearly defined main lobes. The 0.125 and 0.1875 taper F/B optimized antennas have very broad main lobes and are single frequency F/B antennas. Overall, the user in need of a high F/B may find the 0.25 taper F/B optimized antenna to be the best choice.

For the gain-oriented user, the gain optimized antennas with the 0.125 and 0.1875 tapers are a logical choice. Both provide respectable F/B along with a well defined front lobe and an overall clean pattern. While the 0.1875 antenna provides a slight increase in calculated gain, Tilton's 0.125 antenna is every bit as good.

There is a rather intriguing by-product of the Tilton/Greenblum iterations. A boomlength of 3.23 wavelengths is extremely close to the boomlength of 3.2 wavelengths used by Viezbicke. This invites an obvious comparison between the NBS Yagi and the Yagis optimized for this article.

Using the Lawson model to iterate the 3.2 wavelength NBS Yagi results in a computed gain of 15.2

fig. 10. F/B optimized antenna with a 0.25 taper.

dBi.[13] Any of the five Tilton/Greenblum gain optimized Yagis produced gains in excess of this figure. The amount of excess (gain) ranged from 0.132 to 0.398 dB, with a boom only 0.03 wavelengths longer. While longer booms generally result in larger gains, this difference is too small to account for more than the minutest part of the differences in gain. Also, all five gain optimized Yagis have a first minor lobe whose amplitude is from 1 to 4 dB *less* than that of the NBS Yagi. What is even more interesting is the fact that the NBS Yagi uses four more elements than the Tilton/Greenblum Yagi.

The NBS Yagis are element length-optimized with equal director spacing. The Tilton/Greenblum Yagis resulting from computer iteration are also element length-optimized, but director spacing is initially unequal and followed by equally spaced directors. All NBS Yagis use a reflector spacing of 0.2 wavelengths, while the Greenblum design varies reflector spacing as a function of boom length. It would appear that gain optimized Yagis designed as a result of optimizing two variables are more effective than those designed by optimizing a single variable. With fewer elements they are also easier to build. It is only fair to also note that the NBS 3.2 wavelength antenna has an F/B of from 3 to 10 dB above any of the five gain optimized Yagis.

Next month's installment in this series will present computer-iterated alternative Yagis drawn from two well-known 432 MHz design approaches. Iteration-based inferences will be made on a third design, also of long standing. Perhaps as in the case of 220 MHz, 432 MHz may bring a little surprise.

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ham radio

impedance matching:
a brief review

There's no mystery in finding the right match

In April the author discussed the basics of resonant circuits; this month, he walks us through the fundamentals of impedance matching. Both articles are adapted and reprinted with permission from *RF Circuit Design*, published by Howard W. Sams & Company, Indianapolis, Indiana.*

Impedance matching is often necessary in the design of RF circuitry to provide the maximum possible transfer of power between a source and its load. Probably the most vivid example of the need of such a transfer of power occurs in the front-end of any sensitive receiver. Obviously, any *unnecessary* loss in a circuit which is already handling extremely small signal levels simply cannot be tolerated. Therefore, in most instances, extreme care must be taken during the initial design of such a front-end to make sure that each device in the chain is matched to its load.

background

A well-known theorem states that for DC circuits, maximum power will be transferred from a source to its load if the *load resistance* equals the *source resistance*. A simple proof of this theorem is shown in **fig. 1**. In this figure, for convenience, the source is normalized for a resistance of one ohm and a voltage of one volt.

In dealing with AC or time-varying waveforms, however, that same theorem states that the maximum transfer of power from a source to its load occurs when the *load impedance* (Z_L) is equal to the *complex conjugate* of the *source impedance*. *Complex conjugate* simply refers to a complex impedance having the same *real part* with an opposite reactance. Thus, if the source impedance were $Z_S = R + jX$, then its complex conjugate would be $Z_S^* = R - jX$.

If you followed the mathematics in **fig. 1**, then it should be obvious why maximum transfer of power does occur when the load impedance is the complex conjugate of the source. This is shown schematically in **fig. 2**. The source (Z_S), with a series reactive component of $+jX$ (an inductor), is driving its complex conjugate load impedance consisting of a $-jX$ reactance (capacitor) in series with R_L. The $+jX$ component of the source and the $-jX$ component of the load are in series and thus cancel each other, leaving only R_S and R_L which are equal by definition. Since R_S and R_L are equal, maximum power transfer will occur. So when we speak of a source driving its complex conjugate, we are simply referring to a condition in which any *source* reactance is resonated with an equal and opposite *load* reactance, leaving only equal resistor values for the source and the load terminations.

The primary objective in any impedance *matching* scheme then, is to force a load impedance to "look like" the complex conjugate of the source impedance so that maximum power may be transferred to the load. This is shown in **fig. 3** where a load impedance of 2 − j6 ohms is transformed by the impedance

*Available from Ham Radio's Bookstore, Greenville, NH 03048, $24.45 postpaid.

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matching network to a value of 5 + j10 ohms. Therefore, the source "sees" a load impedance of 5 + j10 ohms, which just happens to be its complex conjugate. It should be noted here that because we are dealing with reactances, which are frequency dependent, the perfect impedance match can occur at only one frequency: that is, the frequency at which the +jX component exactly equals the −jX component and thus cancellation or resonance occurs. At all other frequencies removed from the matching frequency, the impedance match becomes progressively worse and eventually non-existent. This can be a problem in broadband circuits where we would ideally like to provide a perfect match everywhere within the broad passband.

There are an infinite number of possible networks which could be used to perform the impedance matching function of **fig. 3**. Something as simple as a 2-element L-C network or as elaborate as a 7-element filter, depending on the application, would work equally well.

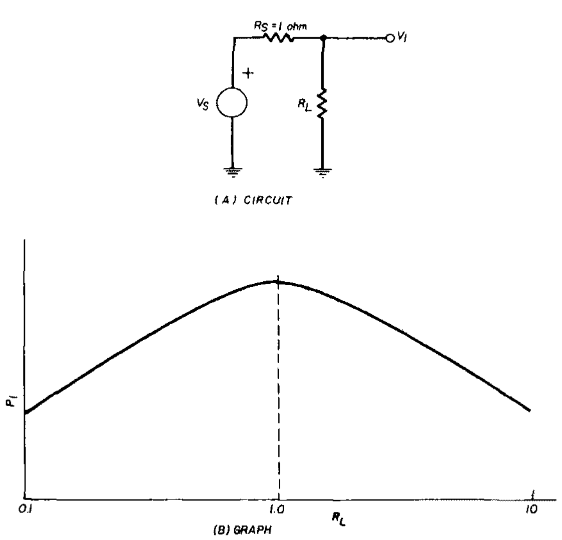

fig. 1. Proof of the power theorem.

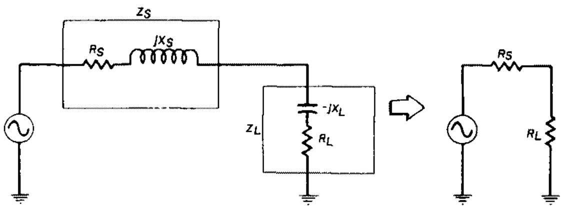

fig. 2. Source impedance driving its complex conjugate and the resulting equivalent circuit.

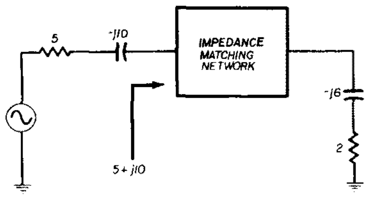

fig. 3. Impedance transformation.

the L-network: why it works

Probably the simplest and most widely used matching circuit is the L-network shown in **fig. 4**. This circuit receives its name from its component orientation, which resembles the shape of an L. As shown in the figure, there are four possible arrangements of the two components. Two of the arrangements, A and B, are in a lowpass filter configuration, while the other two, C and D, are in a highpass filter configuration.

Before we introduce equations that can be used to design the matching networks of **fig. 4**, let's first analyze an existing matching network so that we can understand exactly how the impedance match occurs. Once this analysis is made, impedance matching should seem less mysterious.

Figure 5 shows a simple L-network impedance match between a 100-ohm source and a 1000-ohm load. Without the impedance matching network installed, and with the 100-ohm source driving the 1000-ohm load directly, one-third of the signal *available* from the source is gone before we even get started. The impedance matching network eliminates this loss

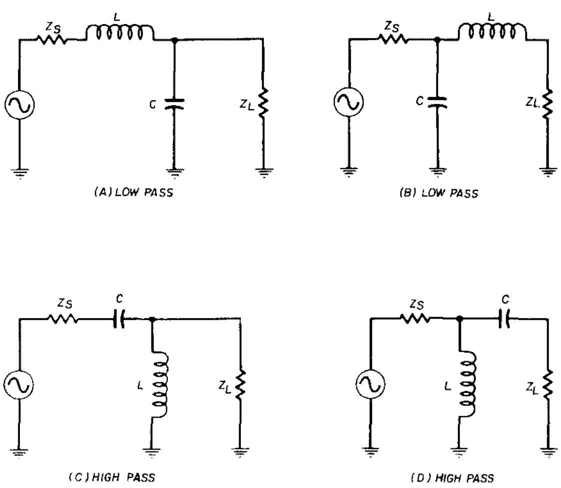

fig. 4. The L-network configured as a low-pass or high-pass circuit.

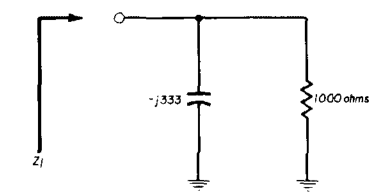

fig. 5. Simple impedance match between a 100-ohm source and a 1000-ohm load.

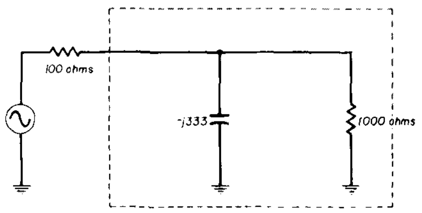

fig. 6. Impedance looking into the parallel combination of R_L and X_c.

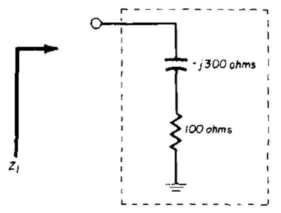

fig. 7. Equivalent circuit of *fig. 6*.

and allows for maximum power transfer to the load. This is done by forcing the 100-ohm source to see 100 ohms when it looks into the impedance matching network. But how?

If you analyze **fig. 5**, the simplicity of how the match occurs will amaze you. Take a look at **fig. 6**. The first step in the analysis is to determine what the load impedance actually looks like when the −j333 ohm capacitor is placed across the 1000-ohm load resistor. This is easily calculated by:

$$Z = \frac{X_c R_L}{X_c + R_L} = \frac{-j333(1000)}{-j333 + 1000}$$

$$= 100 - j300 \text{ ohms}$$

Thus, the parallel combination of the −j333 ohm capacitor and the 1000-ohm resistor *appears* to be an impedance of 100 − j300 ohms. This is a *series* combination of a 100-ohm resistor and a −j300 ohm capacitor as shown in **fig. 7**. Indeed, if you hooked a signal generator up to circuits similar to **figs. 6** and **7** you would not be able to tell the difference between the two as they would exhibit the same characteristics (except at DC, obviously).

Now that we have an *apparent* series 100-j300 ohm impedance for a load, all we have to do to complete the impedance match to the 100-ohm source is to add an equal and opposite (+j300 ohm) reactance in series with the network of **fig. 7**. The addition of the +j300-ohm inductor causes cancellation of the −j300 ohm capacitor, leaving only an *apparent* 100-ohm load resistor. This is shown in **fig. 8**. Keep in mind here that the actual network topology of **fig. 5** has not changed. All we have done is to analyze small portions of the network so that we can understand the function of each component.

To summarize then, the function of the *shunt* component of the impedance matching network is to transform a larger impedance down to a smaller value with a real part equal to the real part of the other terminating impedance (in our case, the 100-ohm source). The series impedance matching element then resonates with or cancels any reactive component present, thus leaving the source driving an apparently equal load for optimum power transfer. So you see, the impedance match isn't mysterious at all; it can be completely explained every step of the way.

Now back to the design of the impedance matching networks of **fig. 4**. These circuits can be very easily designed using the following equations:

$$Q_s = Q_p = \sqrt{\frac{R_p}{R_s} - 1} \tag{1}$$

$$Q_s = \frac{X_s}{R_s} \tag{2}$$

$$Q_p = \frac{R_p}{X_p} \tag{3}$$

where, referring to **fig. 9**:

Q_s = Q of the series leg
Q_p = Q of the shunt leg
R_p = shunt resistance
X_p = shunt reactance
R_s = series resistance
X_s = series reactance

X_p and X_s may be either capacitive or inductive reactance, but each must be of the opposite type. Once X_p is chosen as a capacitor, for example, X_s must be an inductor and vice-versa.

example 1

Design a circuit to match a 100-ohm source to a 1000-ohm load at 100 MHz. Assume that a DC voltage must also be transferred from the source to the load.

Solution: The need for a DC path between the source and load dictates the need for an inductor in the series leg as in **fig. 4A**.

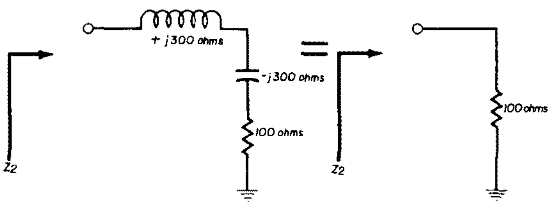

fig. 8. Completing the match.

From eq. 1 we have:

$$Q_s = Q_p = \sqrt{\frac{1000}{100} - 1} = \sqrt{9} = 3$$

From eq. 2:

$$X_s = Q_s R_s = (3)(100) = 300 \text{ ohms (inductive)}$$

From eq. 3:

$$X_p = \frac{R_p}{Q_p} = \frac{1000}{3} = 333 \text{ ohms (capacitive)}$$

The component values at 100 MHz are:

$$L = \frac{X_s}{2\pi f} = \frac{300}{2\pi(100 \times 10^6)} = 477 \text{ nH}$$

$$C = \frac{1}{\omega X_p} = \frac{1}{2\pi(100 \times 10^6)(333)} = 4.8 \text{ pF}$$

This yields the circuit of **fig. 10**. Notice that what you have done is to design the circuit which was previously given in **fig. 5** and analyzed.

dealing with complex loads

The design of **example 1** was for the simple case of matching two *real* impedances (pure resistances). It is very rare when such an occurrence actually exists in the real world. Transistor input and output impedances are almost always *complex*; that is, they contain both resistive and reactive components ($R \pm jX$). Transmission lines, mixers, antennas, and most other sources and loads are no different in that respect. Most will always have some reactive component which must be dealt with. It is, therefore, necessary to know how to handle these stray reactances, and in some instances, to actually put them to work for you.

There are two basic approaches in handling complex impedances as outlined below:

Absorption. It is possible to actually absorb any stray reactances into the impedance matching network itself. This can be done through prudent placement of each matching element such that element capacitors are placed in parallel with stray capacitances, and element inductors are placed in series with any stray inductances. The *stray* component values are then subtracted from the *calculated* element values, leaving new element values, C′,L′, which are smaller than the calculated element values.

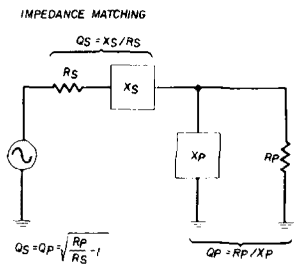

fig. 9. Summary, L-network design.

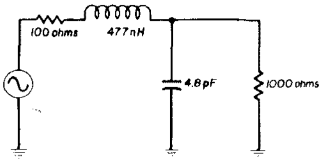

fig. 10. Final circuit of ***example 1***.

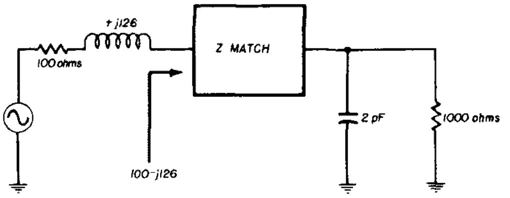

fig. 11. Complex source and load to be matched in ***example 2***.

Resonance. Resonate any stray reactance with an equal and opposite reactance at the frequency of interest. Once this is done, the matching network design can proceed as in **example 1** for two pure resistances.

Of course, it is possible to use both of the approaches outlined above at the same time. In fact, the majority of impedance matching designs probably do utilize a little of both. Let's take a look at two simple examples to help clarify matters.

example 2

Use the absorption approach to match the source and load of **fig. 11** at 100 MHz.

Solution: The first step in the design process is to totally ignore the reactances and simply match the 100-ohm real part of the source to the 1000-ohm real part of the load at 100 MHz. Keep in mind that you would like to use a matching network that will place element inductances in series with stray inductance and element capacitances in parallel with stray capacitances. Conveniently, the network of **fig. 4A** is

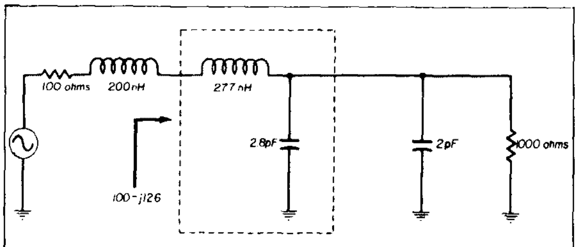

fig. 12. Final design of *example 2.*

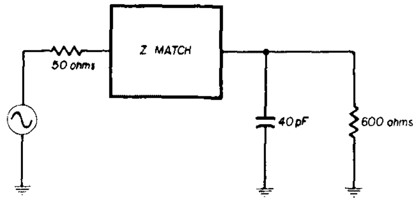

fig. 13. Complex load to be matched in *example 3.*

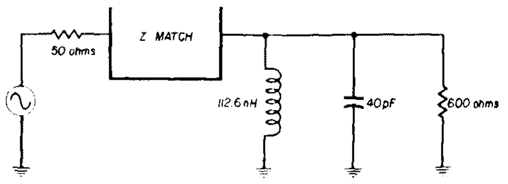

fig. 14. Resonating the stray load capacitance.

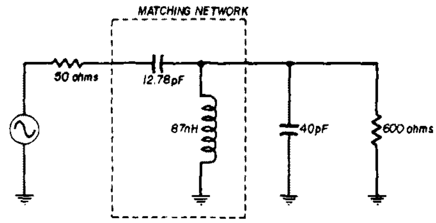

fig. 15. Circuit of *fig. 14* after impedance matching.

again chosen for the design, and again **example 1** provides the details. Thus, the calculated values for the network, if we ignore stray reactances, are shown back in **fig. 10**. But since the stray reactances really do exist, the design is not yet finished as we must now somehow absorb the strays into the matching network. This is done as follows. At the load end we need 4.8 pF for the matching network. We already have a stray 2 pF available at the load, so why not use it? If we use a 2.8 pF *element* capacitor, the *total* shunt capacitance becomes 4.8 pF, the design value. Similarly, at the source, the matching network calls for a series 477 nH inductor. We already have a +j126 ohm, or 200 nH inductor available in the source. If we use an actual element inductance of $477\,nH - 200\,nH = 277\,nH$, then the *total* series inductance will be 477 nH, which is the calculated design value. The final design is shown in **fig. 12**.

Notice that nowhere in the example was a conjugate match even mentioned. However, you can rest assured that if you perform the simple analysis outlined in the previous section of this article, the impedance looking into the matching network, as seen by the source, will be 100 − j126 ohms which is indeed the complex conjugate of 100 + j126 ohms.

Obviously, if the stray element values are larger than the calculated element values, absorption cannot take place. If, for instance, the stray capacitance of **fig. 11** were 20 pF, we could not have added a shunt element capacitor to give us a total needed shunt capacitance of 4.8 pF. In a situation such as this, when absorption is not possible, the concept of resonance coupled with absorption will often do the trick.

example 3

Design an impedance *matching network* which will block the flow of DC from the source to load of **fig. 13**. The frequency of operation is 75 MHz. Try the resonant approach.

Solution: The need to block the flow of DC from the source to the load dictates the use of the matching network of **fig. 4C**. But first let's get rid of the stray 40 pF capacitor by resonating it with a shunt inductor at 75 MHz.

$$L = \frac{1}{\omega^2 C_{stray}}$$

$$= \frac{1}{[2\pi\,(75 \times 10^6)]^2\,(40 \times 10^{-12})}$$

$$= 112.6\ nH$$

This leaves us with the circuit of **fig. 14**. Now that we have eliminated the stray capacitance, we can proceed with the matching network between the 50-ohm load and the apparent 600-ohm load.

Thus:

$$Q_s = Q_p = \sqrt{\frac{R_p}{R_s} - 1} = \sqrt{\frac{600}{50} - 1} = 3.32$$

$$X_s = Q_s R_s = (3.32)(50) = 166\ ohms$$

$$X_p = \frac{R_p}{Q_p} = \frac{600}{3.32} = 181\ ohms$$

Therefore, the element values are:

$$C = \frac{1}{\omega X_s} = \frac{1}{2\pi(75 \times 10^6)(166)} = 12.78\ pF$$

$$L = \frac{X_p}{\omega} = \frac{181}{2\pi(75 \times 10^6)} = 384\ nH$$

These values then, yield the circuit of **fig. 15**. But notice that this circuit can be further simplified by simply replacing the two shunt inductors by a single inductor.

Therefore:

$$L_{new} = \frac{L_1 L_2}{L_1 + L_2} = \frac{(384)(112.6)}{384 + 112.6} = 87\ nH$$

The final design appears in **fig. 16**.

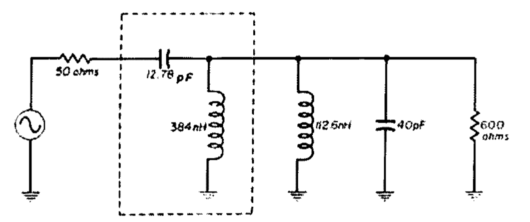

fig. 16. Final design of *example 3*.

conclusion

Examples 2 and **3** detail some very important concepts in the design of impedance matching networks. With a little planning and preparation, the design of simple impedance matching networks between complex loads becomes a simple number-crunching task using elementary algebra. Any stray reactances present in the source and load can usually be absorbed into the matching network, (**example 2**) or can be resonated with an equal and opposite reactance which is then absorbed into the network instead (**example 3**).

Impedance matching isn't really magic at all, is it?

ham radio

ham radio TECHNIQUES

Bill Orr W6SAI

fifty years ago

1934 was a very good year. True, the United States was in the grip of a worldwide depression and unemployment ran over 30 percent, nationwide. But for Radio Amateurs it was a golden year of opportunity. Interest in the hobby was booming and the number of licensed Amateurs grew rapidly. The advent of the AC-operated broadcast receiver and the loosening of patents on vacuum tubes and the superheterodyne receiver brought forth new circuits, new tubes, and new techniques.

For nearly a decade Amateur Radio had been in the doldrums, overshadowed by the "broadcast craze." But now it was coming into its own. For a newly licensed Radio Amateur (like me) the opportunities were numerous. *QST* blazed with enticing advertisements for such newly-created companies as Hallicrafters, Collins Radio Co., and Eitel-McCullough (EIMAC).

Getting on the air, however, was a formidable problem for a high school lad with a weekly allowance of fifty cents. That, plus money earned on odd jobs around the town, soon grew to a grand sum of about fifteen dollars that could be spent on a ham transmitter, and a few months after my license arrived, I was on the air with a 40 watt phone/CW "breadboard" transmitter.

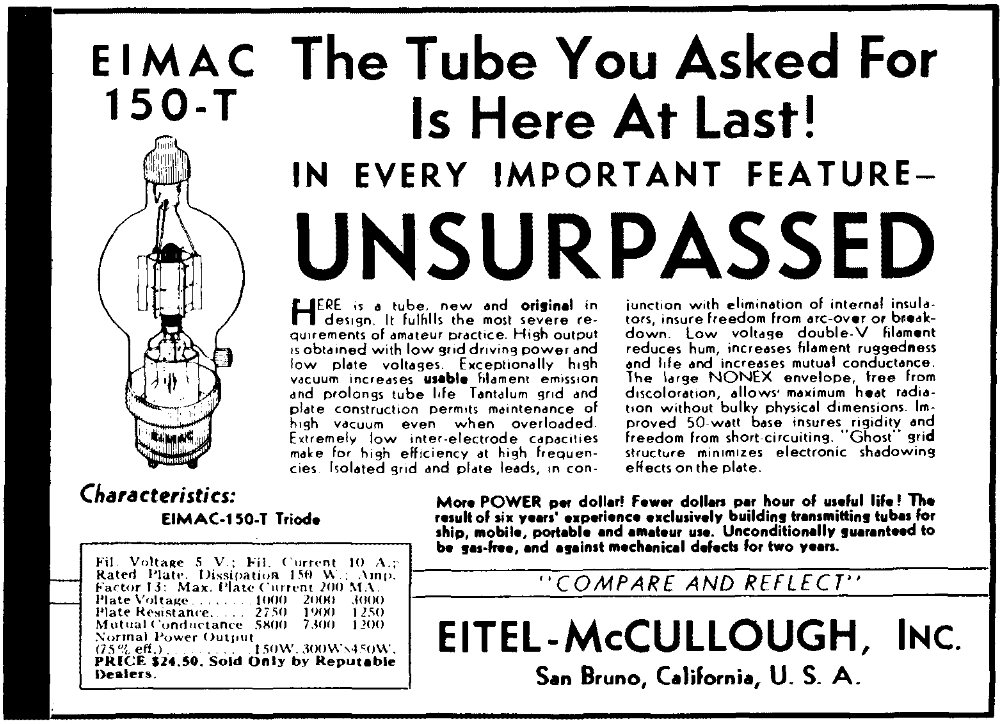

fig. 1. This modest ad started it all in 1934. W6UF and W6CHE couldn't buy the tube they wanted, so they built their own.

It was the beginning of a long journey, still in progress. In 1934, I had little inkling of where the future would lead me.

Across the continent, in San Bruno, California, two young Radio Amateurs were betting their future on a new tube development. Bill Eitel, W6UF, and Jack McCullough, W6CHE, combined their brains, talent, and their thin pocketbooks to start manufacturing a precedent-setting transmitting tube. Little did they realize they would revolutionize the world's tube industry and quickly assume a dominant role in the communications industry. And little did I realize, as I read the first EIMAC ad in November, 1934s *QST* (**fig. 1**), that these two pioneers would become my close friends and colleagues in years to come. Yes, 1934 was a very good year.

more on video disc RFI

In last month's column I commented on the severe RFI problems encountered with the video disc player which has active circuits that function in the 900-925 MHz range. No sooner had I written the column than I received the ARRL Letter which said, in part, that the FCC "had been informed by RCA Corporation that there is a potential interference problem resulting from the new 902-928 MHz secondary status allocation to the Amateur Radio service under Part 2 of the Rules."

RCA went on to state that the FCC should "seek recognition of the need for balancing a possible continuing experimental use of the 902-928 MHz band against the beneficial influence of the video disc player in the lives of millions of U.S. consumers, now, and for years to come."

Amazing! RCA, which has known of the potential incompatibility of its video disc player (as discussed in this column last month), could have chosen to redesign the unit to be immune to RF pickup, but instead chose to lay a "guilt trip" on the Radio Amateur and the FCC. The ARRL promises strong opposition to this move. I'll keep you informed!

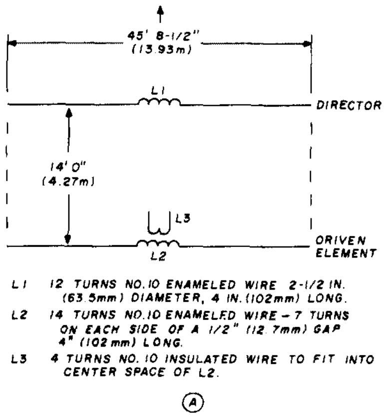

fig. 2A. Top view of 40-meter beam. Dimensions are for 7.2 MHz. Add 10 inches (254 mm) to tips for resonance at 7.025 MHz. Note that both elements are the same length.

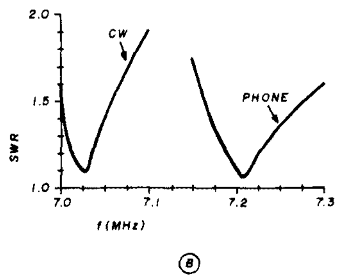

fig. 2B. SWR curves of 40-meter beam for CW and phone operation.

. . . and on RF lamps

Since the late 1970's several companies (General Electric, North American Phillips, International Energy Conservation System, and Soli-Tronics — among others) have been developing and manufacturing limited numbers of RF lighting devices. In general, these are fluorescent tubes that have electronic ballasting instead of an electromagnetic ballast.

One form of RF light is a self-contained lamp that screws into a standard bulb socket; a second type is an external solid-state RF ballast package that will replace the conventional ballast device in existing fluorescent lamps; and a third type of lamp makes use of RF energy delivered to it over the wires.

Most electronic ballasts operate between 20 and 40 kHz, but at least one type operates in the ISM (Industrial-Scientific-Medical) band at 13.56 MHz.

The general theory of operation is that rectified AC is applied to an inverter whose output is RF energy at a frequency above 20 kHz. The RF energy is then applied to a fluorescent tube to strike an arc that excites the fluorescent coating to emit light.

Some of the RF lamps require that the RF oscillator be on continuously while the lamp is lit, while others require a burst of RF energy only when the lamp is turned on.

Compared to an incandescent lamp, the RF lamp is supposed to be more energy-efficient, using only 25 percent as much energy to produce the same lumen output as a conventional bulb.

A powerful argument exists, then, for the marketing of the RF lamp, if only as an energy conserver. The unanswered question is, how much RF interference do these lamps generate? And what will be the effect upon radio communications when hundreds of thousands (or millions) of these lamps are in daily use?

Last fall the FCC granted a limited waiver to various companies to manufacture and market 10,000 electronic ballast units and 100,000 RF light bulbs to be used in field testing and evaluations. The companies will study the cumulative effect of a large number of devices (installed in one plant and all connected to a single wiring system) on the amount of RF interference created, and how this correlates with the RFI level of a single device.

All well and good, but this operation reminds me of a student grading his or her own exam paper! I'm sure these noise generators will soon be on the market; the question is how much control will be exerted by the FCC over the noise radiated by these new RF lamps?

a compact 40-meter beam

As the sunspot cycle continues on its downward trend, activity picks up on the lower frequency bands that are

less affected by the rise and fall of the MUF (Maximum Usable Frequency). Forty meters is *really* coming into its own as a DX band! Some very effective beams are being used on this band, making it very hard for the average ham with a dipole or groundplane to enjoy contacts with exotic DX stations. One answer to this vexing problem is the miniature beam antenna. Despised by those who own full-size beams, the "mini-beam" can give a good account of itself provided it is well designed and properly built. Even a 40-meter mini-beam is quite large, and it's difficult to build one that won't fall apart in heavy wind. Shown in **fig. 2A** is a practical and rugged mini-beam design that has stood the test of time. Used by various California DXers for a decade, it can hold its own in a pile-up and also endure buffeting by heavy winds.

Center-loaded elements are used even though loading coils placed near the element tips are theoretically more efficient. The elements are made of 1 1/4-inch (31.75 mm) OD aluminum tubing with telescoping tips. Twelve foot (3.66 meters) long tubes are used.

The parasitic element is a director, and for CW operation at the low end of the band, is resonant at about 6.7 MHz. The driven element is resonant at 7.025 MHz. The elements are adjusted to resonance with the aid of a dip-oscillator before the beam is assembled. The coils are fixed and frequency adjustments are made to the tip sections.

Loading coils are wound on a 2 1/2 inch (6.35 mm) OD phenolic rod and are given a coat of epoxy after completion. RF current in the coils is quite high, so solid connections must be made between the coil and the elements with 1/2-inch (12.7 mm) wide copper strap.

The elements are supported on a 15-foot (4.57 m) long boom of 2 1/2-inch (63.5 mm) diameter, heavy wall aluminum tubing. Mounting plates and U-bolts hold the elements to the boom. Insulating sleeves are used between the U-bolts and the element sections, as illustrated.

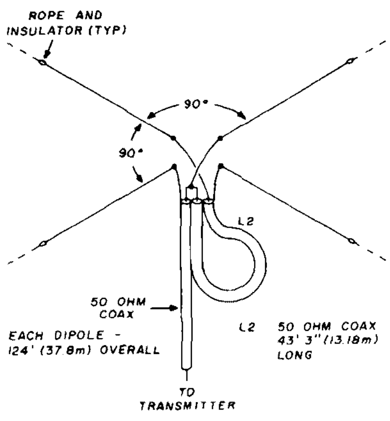

fig. 3. The 80-meter wideband antenna at ZS6ZO. Two dipoles spaced 90 degrees apart in plan view are fed 90 degrees out-of-phase with an electrical quarter-wavelength interconnecting line. Shields of lines are all soldered together at dipole feedpoint and connected to adjacent antenna sections.

SWR curves for the mini-beam are shown in **fig. 2B**, for the CW and SSB band segments. The SWR curve is affected by the director length, primarily, and by the adjustment of the coupling coil at the center of the driven element. Changing the director length by a few inches should drop the SWR curve down to a reasonable value at the design frequency.

Proper operation requires that the beam be well up in the air; a half-wavelength above ground is suggested as a minimum height. That means about 65 feet! Good results have been achieved with the beam as low as 35 feet, but the advantages of low angle radiation are lost when the beam is placed at a low elevation.

the ZS6ZO wideband 80-meter antenna

Dave, ZS6ZO, has had good luck on 80 meters with the Turnstyle-type antenna shown in **fig. 3**. He uses two dipoles cut to the middle of the band and spaced 90 degrees apart. He feeds them 90 degrees out-of-phase with a quarter-wave length of coaxial line between the dipoles. This provides circular polarization to the zenith and omnidirectional horizontal polarization to the horizon.

As one dipole increases in electrical length with respect to the design frequency, the other, via the quarter-wave line, appears shorter. This results in doubling the bandwidth over just two dipoles fed in *parallel* at the center points.

the forgotten RG-58 cable

Save money on your feedline? That's always an attractive proposition. It's not always necessary to use the expensive RG-8A/U or RG-213/U coax, especially for a lower frequency antenna when moderate power levels are used. This is where the RG-58 family of cable comes into use. As with other cables, there are several forms of RG-58 on the market: RG-58/U, RG-58/U type, RG-58A/U, RG-58A/U type and RG-58C/U type. The first two are older style cables with an impedance of 53.5 ohms. Stay away from these, because modern SWR meters are designed for 50-ohm line (the RG-58/U type may also be an inferior cable). The newer, 50-ohm cables are the RG-58A/U and the RG-58C/U. Of the two, the C/U is the better choice because of a non-contaminating (longlife) outer jacket. (The RG-58A/U and the RG-58A/U type both have the lower cost, PVC outer jacket which has a much shorter life.)

The RG-58C/U, when used below approximately 10 MHz, has only about 1 dB loss per one hundred feet, with the loss dropping as the frequency goes down. While the cable is not *rated* in terms of power carrying ability, I have used it with no problems at 1 kW PEP and CW input. Indeed for short runs of up to 25 or 30 feet, it can be used with success up to 30 MHz. At 28.6 MHz a 30 foot section runs slightly warm with 1 kW PEP input working into a load SWR of 1.5:1.

In order to use the cable with the popular PL-259 style of plug an adapter is required. The military number of this part is UG-175/U (Amphenol No. 83-185). The adapter fits inside the

PL-259 and allows the user to make a tight connection between plug and cable.

using the cable adapter

It is tempting to use the cable/plug assembly drawings shown in most handbooks, but I've found a simpler process that allows you to use the plug and adapter more than once. In other words, this method is easier to use and to disassemble than the process outlined in the handbook.

Strip the cable jacket back by 3/4 inch. (I use a sharp nail scissors to do this to prevent nicking the braid.) Trim the end of the jacket square; then, using the scissors, cut the braid back so that only 1/4 inch projects out from under the jacket. At this point (or before) the adapter and PL-259 outer coupling ring are slid over the cable, leaving just the short braid projecting from the rim of the adapter.

Spread the braid out evenly over the rim of the adapter and with the scissors, trim it back to the outer edge of the adapter. Only a fraction of an inch of the braid covers the lip of the adapter now. Next, solder the braid to the lip of the adapter all around the rim. (Use only a small soldering iron to avoid overheating the center insulation of the cable.) When the adapter has cooled, file the rough edges of the braid and solder down to a smooth surface. Thread the adapter and cable into the PL-259 plug and twist the plug/adapter combination tight with the aid of two pliers. Solder the inner conductor of the cable to the end of the center pin of the plug.

It actually takes longer to describe the operation than to do it. The connection has never worked loose in my experience, and it is very easy to unsolder the adapter and reuse it.

Remember that the PL-259 series of plugs are not waterproof; they should be protected against moisture regardless of the assembly technique used. I wrap mine with several layers of electrical tape and that seems to do the job.

ham radio

the ham notebook

vary because your ground situation is unique to your QTH. However, the given coil information should be close. My successes have been "WAS" on 80/40 and 42 on 160 so far. I have worked 66 countries on top band and 74 on 80/40 this year, and also worked four JA stations for WAC on 160 again (last year I worked only one JA). I hope you can install the same kind of system I have had so much success with while "confined" to a townhouse.

Fred C. Race, W8FR

limited space inverted "L"

During the past two years I have been assigned to Great Lakes Naval Training Center and have lived in a townhouse at Grayslake, Illinois. Needless to say, I never thought that I would be able to get on the air from this location at all — let alone on 160, 80, and 40 meters. How did I do it? With a good earth reference and some short inverted "L"s.

The ground reference for the antennas is a "window-well" retaining wall. The dimensions are 4 × 5 feet with 2 feet on each side, leaving a total of 80 square feet of visible ground contact.

The antennas, all "shunt-fed," measure 14 feet in the vertical dimension. I spaced them across the front of the window-well 18 inches from the side of the house. Switching is made possible with a Heathkit antenna switcher. A single line trails into the basement via the air-conditioner pipe hole-through.

Figure 1 illustrates the three-band antenna configuration. Coil taps may

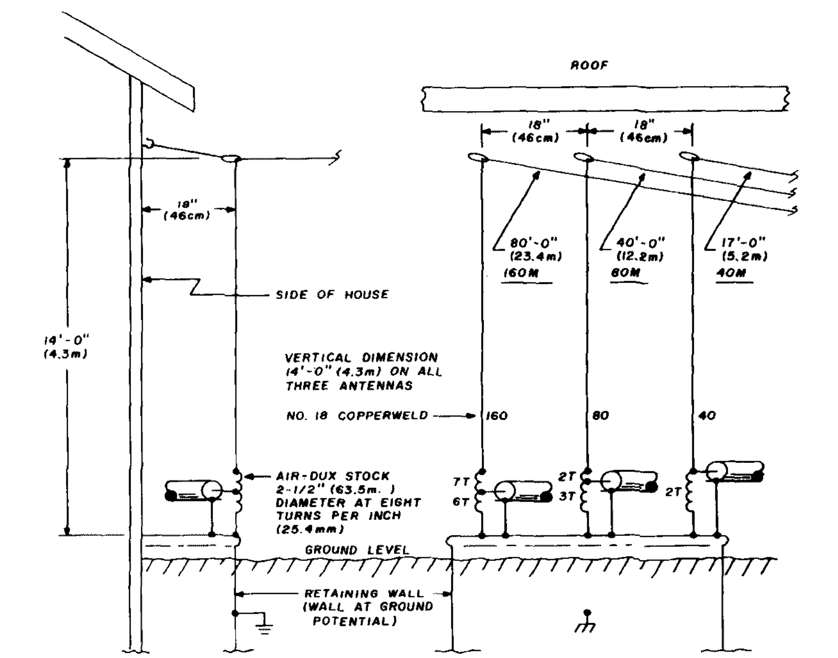

fig. 1. Vertical sections of the shortened inverted L are only 14 feet in length.

neutralizing 572B final at 1500 watts output

The recent FCC ruling setting 1 1/2 kW as the maximum power output prompted me to rework one of my home-made amplifiers, a pair of 572Bs, grounded grid, to grounded cathode for more output.

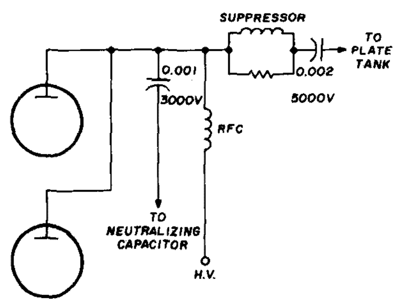

fig. 1. New neutralizing circuit.

The two tubes, fan cooled, would put out 550 watts in grounded grid before showing any color. Grid driven in Class C, they put out a nice 1425 watts. But there was a problem with the frequency-sensitive neutralizing settings. I could adjust the neutralization for stability on any one band, but it would sometimes take off on some other band. In the past I solved this problem by using a form of negative feedback in the filament circuit. However, this requires more drive power on the higher bands, and Class C is hard enough to drive as is.

After much fruitless cut-and-try with

parasitic coils, I decided to take the parasitic coils out of the neutralizing path. I wired the two tubes directly to the hot end of the RF plate choke. From the same point I connected a 0.001 μF 5000 volt mica capacitor to the neutralizing capacitor. Also from the same point I wired in one of the parasitic suppressors to the regular plate coupling capacitor. (See **fig. 1.**) This cured any tendency towards instability. Now with zero bias on the tubes and a full 2750 volts on the plates, no amount of band switching or knob twisting will show the tiniest twitch on the meters.

The parasitic choke is a self-supported coil consisting of seven turns of No. 14 wire, 5/16 inch (8 mm) diameter and 3/4 inch (19 mm) long. The swamping resistor is a Globar with a cold resistance of 400 ohms.

John Labaj, W2YW

high-frequency dummy load

This 52-ohm dummy load consists of twelve 620 ohm, 2-watt resistors housed in a salve can. The load is useful to 175 MHz.

It is capable of dissipating 30 watts on a 50 percent cycle and 50 watts on a shorter duty cycle.

Start construction by filing the outboard shoulder of a UG-176/UHF sleeve to a 3/32 inch (2.5 mm) height. Tin the inner shoulder of the sleeve and outboard end of the PL-259 fitting. The tinned areas are indicated as "solder" in the drawing.

Next, drill and ream a hole in the center of the bottom salve can to accommodate the UG-176/UHF sleeve. Then drill twelve No. 50 drill holes around the periphery of the can. They are on a line half way up the can. Tin area inside and out around each hole, mount the coax connector on the bottom and sweat solder in place.

Solder a 2-3/4 inch (70 mm) length of No. 14 wire center of PL-259 connector extending into the salve can. Fit each 620 ohm 2-watt resistor in place between the center conductor and hole drilled in the rim of the can. Clip

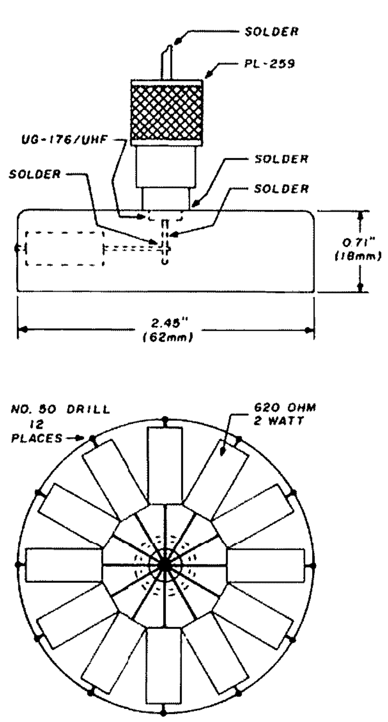

fig. 1. HF dummy load, construction details.

outer end about 1/32 inch (1 mm) beyond can and solder. Form resistor wire approximately half way around the No. 14 wire. After 12 resistors are mounted, consolidate their other ends about the No. 14 wire and solder.

Clip excess from No. 14 wire. Fit on the cover and you have a shielded dummy load.

William J. Goodwin, W1KWE

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antenna support

I have been using nylon rope to hold up my vertical antennas. Nylon rope, however, tends to stretch and come apart, and it needs constant attention.

A better method is to use nylon fishing line. For small vertical antennas, a 300 to 600 pound-test line is okay. For larger antennas ask for cod line, which is almost 1/16-inch thick and very tough. Look in your fishing book for the right knots.

Ed Marriner, W6XM

microphone calibration

Don't guess — here's how to determine your microphone's frequency response with or without a computer

A look around the serious ham's shack will often reveal many pieces of test equipment used to monitor the performance of commercial gear and run tests during the construction of homebrew gear. This equipment is also invaluable for "checking" items obtained during those innumerable scrounging trips that legend says soon become part of every Amateur's life.

Yet, the same serious Amateur who wouldn't dream of trying out a new antenna without first pulling out a trusty VSWR meter may think nothing of sticking a loudspeaker in a box without testing to see whether there are any nasty resonances, and will mount a microphone element in a holder other than the one for which it was designed without doing any tests at all. This lack of quantitative measurement in the performance of acoustic equipment is apparent in the poor audio quality of many stations you hear on the air. In light of the fact that many hams also consider themselves audio buffs, it's surprising that more testing isn't done and that discussions often include such inaccurate descriptions of sound quality as "wide-range," "boomy," "vibrant."

One reason more testing isn't done is that a calibrated microphone is necessary for truly accurate results. Calibrated microphones are expensive and their purchase difficult to justify, considering the few times that they are needed. This article describes a method of calibrating *any* microphone for use as a standard. You don't need an expensive microphone; in fact, instead of a microphone, a small loudspeaker will be sufficient. Any irregularities in the frequency response are not important. The calibration curve will reveal their location and magnitude, thereby enabling the proper allowances to be made.

Those of us who like to use only the latest technology may be disappointed to learn that the method used, the Reciprocity Principle, was described in a book entitled *Theory of Sound* written by Lord Rayleigh in 1877. It apparently wasn't new even at that time.

In applying the principle to microphone calibration, two loudspeakers, (one of which must be reversible), and the microphone under test are employed. When we say that one of the loudspeakers must be reversible, we mean simply that it must be capable of functioning as both a loudspeaker and a microphone. Common cone-type speakers with voice coils meet this requirement.

The first step consists of setting up the equipment shown in **fig. 1**. Loudspeaker SP1, and amplifier A serve simply as a sound source and do not need any special features. The loudspeaker of your stereo system, already connected to an amplifier, can be used as a convenient SP1.

Loudspeaker SP2, which must be reversible, and microphone M, whose calibration we seek, are placed side by side at distance *d* in front of SP1. (Distance *d* should be the same as the distance you intend to use for future tests because the low-frequency response of many microphones varies slightly with distance from the sound source.) The distance selected will depend upon the nature of the information desired; a microphone-to-speaker spacing of less than two feet greatly excludes the acoustic effects of

By Daniel Peters, NY6U, Falcon Communications, P.O. Box 620625, Woodside, California 94062

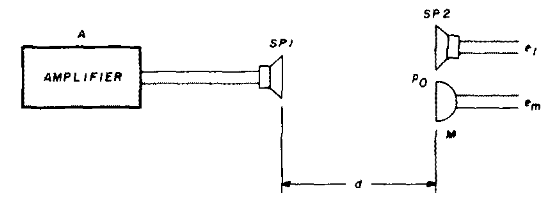


fig. 1. First step toward microphone calibration requires reversible loudspeaker.

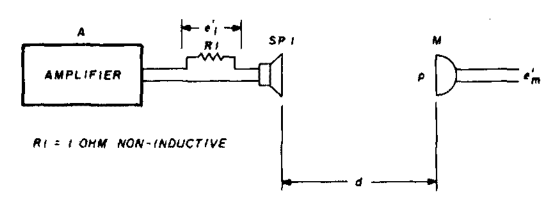


fig. 2. Insertion of one ohm resistor in amplifier output provides convenient measurement terminals.

the room, while larger spacings provide a measure of the effect of room acoustics. If you are calibrating your usual station microphone, it would be well to set distance *d* the same as your normal mouth-to-microphone distance.

Because the high frequency response of most microphones changes for different angles of incidence, position M and SP2 so that they point directly at SP1. For the same reason, place M and SP2 close to each other and aim SP1 squarely at them.

Feeding an audio tone into SP1 results in a sound pressure P_o at M and SP2, which generates an open circuit voltage e_m and e_l, respectively. Measure and record these voltages at each frequency of interest, using either a millivoltmeter or an audio amplifier followed by an AC voltmeter. As long as you use the same meter to make all measurements, the frequency response of the measuring instrument will not affect the results. Likewise, it is not necessary that SP1, or the audio signal driving it, have a flat, or even known amplitude. If you do not have a variable frequency *audio generator* to drive the amplifier connected to SP1, a tape with recorded audio tones will serve. Again, frequency response is not important; all you need is a source of known frequency.

Use a signal into SP1 sufficient to mask any background noise. However, keep the amplitude low enough to prevent overloading; otherwise the harmonics generated will give erroneous results.

Next, connect the equipment as shown in **fig. 2**. Note that we have replaced SP1 with SP2 and have added a 1-ohm resistor in series with the amplifier output. Distance *d* should remain unchanged. Supply audio power and measure voltages e_m' and e_l' at the same frequencies and with the same meter as used in the preceeding steps. Actually the current in SP2 is the quantity of interest; the 1-ohm resistor, R1, provides a 1-volt drop for every ampere of current, thereby allowing the use of the same voltmeter used in our other measurements.

After measuring e_m' and e_l' at the same frequencies at which e_m and e_l were measured, you should end up with a set of data such as shown in table 1.

table 1. Sample microphone calibration data.

f(MHz)	e_1*	e_m	e_1'	e_m'
20	0.50	0.40	190	3.6
30	0.50	0.70	250	6.0
40	0.60	1.50	320	4.8
60	0.40	2.30	320	1.6
80	0.62	2.60	320	5.2
100	2.40	1.40	320	1.8
150	0.65	0.50	320	5.2
200	0.36	1.40	320	5.1
300	0.36	2.70	320	8.8
500	0.74	5.40	320	25.5
700	1.90	7.80	320	170.0
1000	1.50	36.00	320	30.0
1500	2.20	12.00	320	37.0
2000	2.90	14.00	320	84.0
3000	0.70	0.16	320	34.0
4000	0.28	8.20	320	110.0
5000	0.14	6.80	320	54.0
7000	0.08	1.60	320	52.0

*All voltages in millivolts.

The formula for calculating the microphone sensitivity, S_m, is:

$$S_m = K\sqrt{\frac{e_m e_m'}{e_l e_l' f}} \qquad (1)$$

The derivation of the above formula, including the value of *k*, is provided in the appendix. However, since a relative response curve is all that is generally required, the value of *k* is not important and can be left out of the formula, resulting in:

$$S_{mr} = \sqrt{\frac{e_m e_m'}{e_l e_l' f}} \qquad (2)$$

In the above formula, S_{mr} is the relative microphone sensitivity, and *f* is the frequency in Hz; voltages are expressed in volts. If you took your readings in mV, you can use mV, if you use mV for all four entries.

Substitute the values recorded for each frequency in the formula and calculate the relative sensitivities. After finding the relative sensitivity for frequencies of interest, select a value considered an average and calculate all other points in reference to the selected one,

table 2. Sample relative microphone sensitivity calibration chart.

f(MHz)	e_1*	e_m	e_1'	e_m'	S_{mr}	%S_{mr}**
20	0.50	0.40	190	3.6	0.0275	99.21
30	0.50	0.70	250	6.0	0.0334	120.60
40	0.60	1.50	320	4.8	0.0306	110.30
60	0.40	2.30	320	1.6	0.0218	78.89
80	0.62	2.60	320	5.2	0.0291	105.10
100	2.40	1.40	320	1.8	0.0057	20.64
150	0.65	0.50	320	5.2	0.0091	32.89
200	0.36	1.40	320	5.1	0.0176	63.44
300	0.36	2.70	320	8.8	0.0262	94.49
500	0.74	5.40	320	25.5	0.0341	122.90
700	1.90	7.80	320	170.0	0.0558	201.10
1000	1.50	36.00	320	30.0	0.0474	170.90
1500	2.20	12.00	320	37.0	0.0205	73.89
2000	2.90	14.00	320	84.0	0.0251	90.71
3000	0.70	0.16	320	34.0	0.0028	10.25
4000	0.28	8.20	320	110.0	0.0501	180.80
5000	0.14	6.80	320	54.0	0.0404	145.90
7000	0.08	1.60	320	52.0	0.0215	77.65
					average sum 0.0277	

*All voltages in millivolts.

$$**\%\ S_{mr} = \frac{S_{mr}}{S_{mr}(average)} \times 100\%$$

where S_{mr} average = sum of all S_{mr} readings divided by 18 (number of lines of data)

either as a percentage or in dB, using the selected value as 0 dB. (See **table 2.**) (Results may be plotted in the form of a graph.)

You are now ready to use your newly calibrated microphone. One word of caution: when you use the calibrated microphone to test a favorite loudspeaker or microphone that you thought to be "flat," don't be surprised if the response curve resembles a Rocky Mountain skyline.

Individually tuning the ports seen on many microphone cases and loudspeaker enclosures by partially covering or otherwise impeding the air flow with various fabrics can often do much to improve their response. However, even when you've adjusted the system for optimum performance the curve is still going to look pretty rough, and it's better not to mention it to your uninitated friends because they'll inevitably insist that their similar component is "flat." It isn't — and you know it — but why lose friends?

appendix

derivation of formula

Referring to **fig. 1**:

$$e_l = S_l P_o \tag{1}$$

$$e_m = S_m P_o \tag{2}$$

where: S_l = sensitivity of SP2 (in abvolts per dyne per square centimeter)

S_m = sensitivity of M

If we assume SP2 is a conventional loudspeaker, we know that e_l is produced by a conductor moving in a magnetic field, or:

$$e_l = blv \tag{3}$$

where: b = flux density in speaker gap, in Gauss

l = length of wire in the voice coil, in cm

v = velocity of the coil, in cm per second

From the "Ohms law" of mechanical circuits:

$$V = \frac{P_o A}{Z_m} \tag{4}$$

where: A = diaphragm area in square centimeters

Z_m = mechanical impedance of the vibrating system in mechanical ohms

Combining **eqs. 1, 3,** and **4**:

$$S_l = \frac{e_l}{P_o} = \frac{blA}{Z_m} \tag{5}$$

Referring now to **fig. 2**, the pressure p at M, caused by SP2 located d centimeters away, is given by:

$$P = \frac{r_a V}{2\lambda d} \tag{6}$$

where: r_a = acoustic impedance of the atmosphere in mechanical ohms (41.5 for standard air at sea level)

λ = wavelength of sound, in cm

The velocity v for a current of i abamperes in SP2 is given by:

$$v = \frac{bli}{Z_m} \tag{7}$$

Combining **eqs. 6** and **7**:

$$P = \frac{rblAi}{2\lambda d Z_m} \tag{8}$$

From **eqs. 8** and **5**:

$$P = \frac{riSl}{2d\lambda} \tag{9}$$

The pressure p on M produces a voltage e_m' given by:

$$e_m' = S_m P \tag{10}$$

Combining this with eq. **9**:

$$Sm = \frac{2d\lambda e_m'}{S_l ri} \quad (11)$$

Combining eqs. **1** and **2** to eliminate P_o and then substituting for S_1:

$$S_m = \sqrt{\frac{2d\lambda e_m e_m'}{e_l ri}} \quad (12)$$

However, since $i = e_l'$, by virtue of the 1-ohm resistor:

$$S_m = \sqrt{\frac{2d\lambda}{r} \bullet \frac{e_m e_m'}{e_l e_l'}} \quad (13)$$

Finally, remembering that $\lambda = v/f$, where v = velocity of sound in cm/sec:

$$S_m = \sqrt{\frac{2dv}{r}} \bullet \sqrt{\frac{e_m e_m'}{e_l e_l' f}}$$

$$= k\sqrt{\frac{e_m e_m'}{e_l e_l' f}} \quad (14)$$

Thus, using only equipment likely to be found in any well equipped shack, an absolute calibration curve for the sensitivity of a microphone can be obtained without the use of a previously calibrated standard.

microphone calibration program

By Nick Corcodilos, 765 San Antonio Road #51, Palo Alto, California 94303

The program listing provided here has been designed to automate the number-crunching necessary in calibrating your microphone. Although developed for the Radio Shack TRS-80 Model 1® computer, the program should run on other TRS-80s, and, with a little modification, on most computers that have a BASIC interpreter. Even though BASIC is widely used, there is no standard version usable on all machines. So if you're using a Commodore or a Timex/Sinclair, for example, some of the code will have to be modified.

The version of BASIC used here is what Radio Shack calls Level II BASIC. It requires no disk drives. An Epson MX-80 dot matrix printer was used to test the program; other printers should work equally well.

Once you've typed the program into your computer, you'll want to save it for later use. You can do this using disks or a cassette storage device. Look up the "SAVE" command in your computer manual for instructions. The "LOAD" and "RUN" commands will also be useful.

The TRS-80 Model 1 CRT Monitor, (16 lines × 60 characters) can display fourteen calibration samples before it scrolls. The program takes these limits into account. If you have a TRS-80 with a 24 line by 80 character screen, you'll have to modify the program to take full advantage of your monitor.

To start, begin typing in the program. Type in the lines of code exactly as they are listed. The spaces between words are as critical as characters; count the spaces carefully and put them in the right places. There are important differences between commas (,), semicolons (;), and colons (:). Type patiently and check and recheck your typing.

Because the program was written to be understandable, plenty of "Remark" statements are included. These remarks are preceded by a single quote (') either after a line number or at the end of a line of code, and are not executed during a program run. They are included to break up the code and help you see which sections of code do what. When you type these, be sure to include the leading (').

When you run the program, it will prompt you to enter five data values for each sample: the frequency (Hz) of the sample, loudspeaker voltages E(L) and E(L)1 where 1 signifies "prime," and microphone voltages E(M) and E(M)1. (These correspond to e_l,e_l', e_m, and e' in the tables.) Since all values for one sample are to be entered together, you'll have to record your original data on paper while you're taking measurements with L2 in its two different positions. (Note L1 = SP1, L2 = SP2.)

The program will produce three tables at the end of its run. Together, these tables will be roughly equivalent to **table 2** in the article. When each table appears on the screen, you'll be asked if you want to print that table. When responding to the "<P> to print to printer?" prompt, be sure to use a capital "P." Lower case won't work.

The main values you'll be interested in are the Relative Microphone Sensitivities (Smr's) for each sample frequency. Also provided is the Average Smr, which you may or may not be interested in. Keep in mind that the Average Smr is just an intermediate value; it is provided in case you should want to do something with it.

In the interest of simplicity, the program has limited error-recovery capability. BASIC does not allow dividing a number by zero. Because some of the calculations the program performs are divisions (in lines 510 and 598), you may experience this problem. If you enter a zero data value, you'll get a "divide by 0" or "/0" error on the screen and your results will be invalidated; the program will "crash." If you enter a character instead of a number (the program accepts numerical data only), you'll get a "?redo from start" error message. Just retype your data value, using a number this time.

If you enter a value incorrectly (for example, 125 instead of 12.5), you will have an opportunity to correct your error after all values for that one sample have been entered (you cannot correct a zero entry). In other words, you can cancel a sample and re-enter it if you do so before going on to the next sample. You cannot change a sample after all samples have been entered. When responding to the prompt "<X> to correct this sample" be sure to use a capital, not lower case, "X."

Be sure your printer is on before you request the printing of a table. Programs run on the Model 1 computer have been known to "crash" if a printer wasn't ready and waiting when needed.

To cancel the program at any point, press <BREAK>. Type "run" followed by a carriage return to start the program again. When the last table has been displayed on the screen, the program will recycle to its beginning, with all data wiped out. You'll know you're there when you see "MICROPHONE CALIBRATION — NEW DATA" at the top of the screen. New data must then be typed in.

Those who need more than fourteen samples in your calibration efforts can either run the program more than once with the additional samples, or tweak the program a bit; with tweaking, your tables will scroll off the screen because they'll be too big, but if you have a printer, the full tables can be printed there with no losses. To tweak the program, change every "14" in lines 220, 350, and 355 to the number of samples you wish to use. That's it. (If your program crashes after these modifications, it's probably because your computer doesn't have enough memory to handle the new number of samples.)

Note: This program is also available in a version designed for use on the IBM PC. For a copy of that program, send a business-sized SASE to N.A. Corcodilos, 765 San Antonio Road #51, Palo Alto, California 94303. — Editor

```
50 '                  PROGRAM BY N A CORCODILOS 2/25/84
55 '                  DEVELOPED AROUND AN IDEA BY D PETERS
60 '
100 '                 MICROPHONE CALIBRATION PROGRAM - TRS80 MODEL 1
110 '                                                  LEVEL II BASIC
150 '
200 '-------------------------DEFINITION OF VARIABLES
210 '
220 DIM M(14), M1(14), L(14), L1(14), F(14), SMR(14)
250 '
300 '-------------------------DATA ENTRY
310 '
320 CLS
330 PRINT "MICROPHONE CALIBRATION DATA - NEW DATA"
340 PRINT
350 INPUT "HOW MANY SAMPLES WILL YOU ENTER (14 MAX)     ";S
355 IF S<1 OR S>14 THEN 300  'CHECKS FOR TOO MANY OR FEW SAMPLES
400 CLS
430 '
440 FOR J=1 TO S
445   PRINT "MICROPHONE CALIBRATION DATA     ";S;" SAMPLES"
446   PRINT
450   PRINT "SAMPLE # ";J
460   INPUT "F(HZ)        ";F(J)
470   INPUT "E(L)         ";L(J)
480   INPUT "E(M)         ";M(J)
490   INPUT "E(L)1        ";L1(J)
500   INPUT "E(M)1        ";M1(J)
510   I=((M(J)*M1(J))/(L(J)*L1(J)*F(J)))   'INTERMEDIATE VALUE
520   SMR(J)=SQR(I)      'SQR IS A BASIC FUNCTION WHICH TAKES SQ ROOT
530   PRINT
540   PRINT "SMR FOR ";F(J);"HZ = ";:PRINT USING "##.####";SMR(J)
550   PRINT
560   INPUT "<X> TO CORRECT THIS SAMPLE, <ENTER> TO CONTINUE ";Q$
570   IF Q$="X" THEN CLS:PRINT "MICROPHONE CALIBRATION DATA     ";S;" SAMPLES"
575   IF Q$="X" THEN PRINT:Q$="":GOTO 450
580   IF Q$<>"" THEN Q$="":GOTO 560
585   Q$="":CLS
590   NEXT J
591 '
592 '-----------------------------SUM & AVERAGE SMR'S
593 '
594 FOR J=1 TO S
595   H=H+SMR(J)         'H IS A HOLDING VARIABLE
596   NEXT J
597 '
```

```
598 AVGSMR=(H/S)
600 '
610 '-----------------------------PRINT CALIB TABLE 1 TO SCREEN
620 '
630 PRINT "F(HZ)","E(L)","E(M)","    TABLE 1"
640 FOR J=1 TO S
650   PRINT F(J),L(J),M(J)
660   NEXT J
675 INPUT "<P> TO PRINT TABLE TO PRINTER, <ENTER> TO CONTINUE INSTEAD ";Q$
676 IF Q$="P" THEN GOSUB 2000 ELSE IF Q$<>"" THEN 675
678 '
680 '-----------------------------PRINT CALIB TABLE 2 TO SCREEN
681 '
682 CLS:PRINT "F(HZ)","E(L)1","E(M)1"
684 FOR J=1 TO S
685   PRINT F(J),L1(J),M1(J)
686   NEXT J
688 INPUT "<P> TO PRINT TABLE TO PRINTER, <ENTER> TO CONTINUE INSTEAD ";Q$
689 IF Q$="P" THEN GOSUB 3000 ELSE IF Q$<>"" THEN 688
690 '
900 '-----------------------------PRINT CALIB TABLE 3 TO SCREEN
910 '
920 CLS
970 PRINT "F(HZ)","SMR","AVERAGE SMR = ";:PRINT USING "##.####";AVGSMR
990 FOR J=1 TO S
1000   PRINT F(J),
1010   PRINT USING "##.####";SMR(J)
1050   NEXT J
1110 INPUT "<P> TO PRINT TO PRINTER, <ENTER> TO RESTART PROGRAM ";Q$
1115 IF Q$="P" THEN GOSUB 4000 ELSE IF Q$<>"" THEN 1110
1200 RUN         'RESTART PROGRAM AFTER FINISH PROCESSING
1900 '
2000 '---------------------------LPRINT ROUTINE TABLE 1
2020 '
2100 CLS:INPUT "TURN PRINTER ON & ALIGN PAPER - PRESS <ENTER> ";Q$
2110 LPRINT:LPRINT "MICROPHONE CALIBRATION TABLE 1"
2120 LPRINT:LPRINT
2130 LPRINT "F(HZ)","E(L)","E(M)"
2140 LPRINT
2150 FOR J=1 TO S
2160   LPRINT F(J),L(J),M(J)
2170   NEXT J
2175 LPRINT:LPRINT:LPRINT:LPRINT
2178 Q$=""
2180 RETURN
2190 '
3000 '---------------------------LPRINT ROUTINE TABLE 2
3020 '
3100 CLS:INPUT "TURN PRINTER ON & ALIGN PAPER - PRESS <ENTER> ";Q$
3110 LPRINT:LPRINT "MICROPHONE CALIBRATION TABLE 2"
3120 LPRINT:LPRINT
3130 LPRINT "F(HZ)","E(L)1","E(M)1"
3140 LPRINT
3150 FOR J=1 TO S
3160   LPRINT F(J),L1(J),M1(J)
3170   NEXT J
3175 LPRINT:LPRINT:LPRINT:LPRINT
3178 Q$=""
3180 RETURN
3190 '
4000 '---------------------------LPRINT ROUTINE TABLE 3
4010 '
4100 CLS:INPUT "TURN PRINTER ON & ALIGN PAPER - PRESS <ENTER> ";Q$
4110 LPRINT:LPRINT "MICROPHONE CALIBRATION TABLE 3"
4120 LPRINT:LPRINT
4130 LPRINT "F(HZ)","SMR","AVERAGE SMR = ";:LPRINT USING "##.####";AVGSMR
4140 LPRINT
4150 FOR J=1 TO S
4160   LPRINT F(J),:LPRINT USING "##.####";SMR(J)
4170   NEXT J
4175 LPRINT:LPRINT:LPRINT:LPRINT
4178 Q$=""
4180 RETURN
```

ham radio

Joe Reisert W1JR

improving meteor scatter communications

It has been estimated that each day 50 to 100 million particles orbiting in space enter the Earth's gravitational field and are literally swept into the ionosphere. These particles, or meteors, as they are often called, are usually quite small — even as small as a grain of sand — and may leave a bright trail as they burn up by frictional heating when entering the atmosphere.

As meteors burn they produce ionization in the "E" layer (typically 50 to 150 km above the Earth). For many years VHF/UHFers have been using this ionization as one of the primary modes of propagation for working DX on 6 and 2 meters. A form of forward scatter, this is usually referred to as "meteor scatter."

During the late 1950s the race for the first 2-meter WAS began, and meteor scatter became one of the prime propagation modes. It soon became obvious that it was very difficult to work new states, especially when they were over 1200 miles (1930 km) distant. Predicting the best time and day were mainly based on luck or past experience.

When the competition for the first 220 MHz and 70 cm (432 MHz) meteor scatter QSOs began, the difficulty increased even further.[1] In fact, as of today, only about a half-dozen 70 cm QSOs have been claimed since the original one on August 12, 1972.[2]

Over the past ten years I have been gathering information on meteor scatter communications and developing techniques, some already known, and some that are perhaps new to Amateurs. This information may assist you in pinpointing meteor scatter maximums, choosing the optimum dates and the best time to schedule, optimizing your equipment, and learning how to listen for meteors. Understanding this material should increase your success rate and possibly add to the data base that will be presented here. Since the prime meteor scatter season is just around the corner, I thought this would be a good time to present this material.

background information

There are two basic types of meteors: sporadic and shower. The sporadic meteors have random orbits in space and are drawn in by the gravitational force of the Earth as it passes them in its orbit around the sun. However, they are for the most part concentrated toward the ecliptic plane (the plane of the Earth's orbit) and move around the sun in the same direction as the Earth. Because of the geometry involved, these sporadic meteors tend to peak for east-west paths every day around 0600 and are at a minimum at 1800 local time with approximately 4:1 ratio.[3] North-south paths tend to peak around 0200 and 1000 local time with a similar minimum. Sporadic meteors are not uniformly distributed and tend to be random in speed (more on this later) as well as difficult to predict. There is a further seasonal variation because of the tilting of the Earth's axis relative to the ecliptic plane. Sporadic meteors tend to peak with a broad maximum occurrence in July and a broad minimum in February.

Shower meteors, on the other hand, are more spectacular, but account for only a small fraction (less than 5 percent) of the total incidence of meteors. They are believed to be the remnants of old or extinct comets which have specific orbits and velocities around the sun. When the Earth intersects one of these orbits, there is a dramatic increase in the quantity and size of the meteors entering the ionosphere. Hence the name "meteor shower" has been coined to describe this phenomenon. The most observed meteor showers have been given astronomical names corresponding to the constellations in the sky from which they appear to originate. Because of their distinct orbit (in comparison to the sporadic types), they can peak at any time of the day or night. The duration of the shower may last anywhere from a few hours to as long as a week or two. Since they are so concentrated (compared to the sporadic types), they considerably enhance the possibility of a completed QSO.

table 1. Data for major meteor showers.

shower name	E.L.*	best dates	duration	accuracy	hourly rate**	velocity (km/sec)	local*** rise/set
Quadrantids	282.83	Jan 1 - 4	10 hours	± 15 min	50	43	2300-1800
April Lyrids	31.40	April 20 - 23	2 days	± 12 hrs	12	51	2100-1100
Eta Aquarids	44.00	May 2 - 6	5 days	± 12 hrs	15	64	0300-1200
Arietids	75.00	June 1 - 15	8 days	± 12 hrs	66	39	0330-1530
June Lyrids	84.00	June 10 - 21	2 days	± 12 hrs	10	51	2100-1100
Delta Aquarids	125.00	July 26 - 30	2 days	± 12 hrs	20	43	2200-0600
Perseids	139.30	Aug 10 - 14	4 days	± 75 min	49	60	(note 1)
Orionids	207.00	Oct 18 - 23	2 days	± 12 hrs	18	67	2230-0930
Taurids	220.00	Oct 30 - Nov 10	20 days	± 12 hrs	10	31	1900-0630
Leonids	234.70	Nov 14 - 19	3 hours	± 12 hrs	10	72	0000-1230
Geminids	261.20	Dec 10 - 15	3 days	± 12 hrs	60	37	1900-0900
Ursids	270.00	Dec 21 - 24	12 hours	± 12 hrs	15	35	(note 2)

* Ecliptic longitude in 1950 coordinates.
** Estimated meteors per hour at maximum. Can vary greatly from year to year depending on shower.
***For northern mid-latitudes.
Note 1. Never sets. Minimum at 1730.
Note 2. Never sets. Minimum at 2030.

Sporadic meteors will continue to be used by VHFers on 6 and 2 meters. The best times for use will be in the morning hours and during the summer months as stated earlier. However, long haul 2-meter DX and especially operation on 220 MHz and above would best be served by concentrating on specific meteor showers.

pinpointing meteor shower peaks

Although many articles have been written on meteor showers, there has been very little information on pinpointing exactly when these meteor showers peak. Most of the available information lists only the approximate dates of the expected peak.[4,5,6] Some of these showers are of extremely short duration (one to four hours). Considering operation on only one particular day, or even during any six to ten hours off peak on a short duration shower will probably prove to be a waste of time.

Often you hear someone say, "Oh, that shower always peaks at 8 AM on August 12," only to hear conflicting stories about the same shower from someone else. The reasons for contradiction are many. Often overlooked is the basic fact that *the Earth takes 365¼ days to complete one orbit around the sun*. (This is the reason we need one leap year every four years in order to get the calendar back in synchronization.) Therefore the shower we encounter today will typically peak six hours later next year, allowing, of course, for leap year when it occurs. Astronomers have a way to predict the time that the meteor showers are expected to peak.[7] This method uses celestial information to predict the time the Earth intersects the orbit of the meteor stream based on data that has been generated by long-term visual observation of the more well known showers. Some information has also been generated from radar observations. I gave a talk on this prediction method at the Central States VHF Conference in Sioux Falls, South Dakota in July, 1981 and have been continuously updating the handout information that accompanied that presentation. (Each month I use this method to forecast the various meteor shower peaks and list them in the "Important VHF/UHF Events" at the end of the column.)

Let's see how to use this method. First you have to know the ecliptic (sometimes called solar) longitude of the various showers. This is listed each year in various publications[8] and I believe *Sky and Telescope Magazine* uses the same information for its monthly meteor shower predictions. However, *beware of meteor shower peak predictions in astronomy magazines;* their interest is primarily visual. If, for instance, the peak of the shower is during the day or on an evening near full moon, astronomy magazines may not provide sufficient information to determine the real peak. Most astronomers couldn't care less about the use of meteor showers for radio communications!

I have generated **table 1** to show the ecliptic longitude in 1950 (astronomical) coordinates for the peak of the meteor showers, the range of dates and times, the duration of the shower, the approximate accuracy of the predictions, and the hourly rate and velocity. To predict the time of the actual meteor shower peak, you need to acquire a table showing "Ecliptic Longitude at 0000 UTC" for the year of interest[9] or calculate this data yourself (more on this later). Because four years of data are required, this would be an extensive table. Therefore, using reference 9, I have calculated and listed in **table 2** the ecliptic longitude for just the principal dates surrounding the major meteor showers. Days on either side of those listed could be estimated if desired. Because the Earth returns to approximately the same place in its orbit every four years (as described earlier) all you have to do is repeat the proper year. For example, the 1984 table is also good for 1976, 1980, 1988, and so on.

The equation for calculating the ecliptic longitude daily is found in reference 9. W4WD has taken this equation and some of the data shown in **table 1** and written a computer program for the TRS-80 computer.[10] Jim Reisert, AD1C, has revised the program and data. A copy of his program is shown in **fig. 1**.

using the tables

A few examples of how to use **tables 1** and **2** should clarify the method.

Find the date and peak time for the Perseids meteor shower in 1984. Table 1 shows the peak at ecliptic longitude 139.3. Scan through the ecliptic longitudes listed on **table 2** for 1984 and find the day before and the day after the 139.3 peak, noting the date and ecliptic longitude shown for 0000 UTC. The day before is August 11 at ecliptic longitude 138.51, and the day after is August 12 at 139.47. Therefore, the shower will peak on August 11. To find the peak shower time, insert this ecliptic longitude data into the following equation:

$$T = 24 \cdot \frac{(E.L. - E.L.1)}{(E.L.2 - E.L.1)}$$

$$= 24 \cdot \frac{(139.30 - 138.51)}{(139.47 - 138.51)}$$

$$= 24 \cdot (0.823) = 19.75$$

or 1945 hours

where T is time in UTC, $E.L.$ is ecliptic longitude from **table 1** for the shower of interest, $E.L.1$ is from **table 2** for the day before and $E.L.2$ is from **table 2** for the day after peak for the proper year. Therefore, the next peak of the Perseids meteor shower should be August 11, 1984, at approximately 1945 UTC.

Find the peak for the great Leonids shower of 1966. Table 1 shows the peak at ecliptic longitude 234.7. Looking through **table 2** for 1982 (same as 1966 table, as explained earlier) we find the data for the day before and after as November 17 at 234.28 and November 18 at 235.39. Inserting these eclip-

fig. 1. Meteor shower peak prediction time program. Though written for TRS-80™, BASIC program can be adapted to any personal computer.

```
100 REM:  PROGRAM BY JAMES REISERT, AD1C - 12 FEBRUARY 1984
105 REM:  BASED ON PROGRAM BY RUSS WICKER, W4WD
110 K=57.2957795130S
115 PRINT
120 PRINT TAB(5);"METEOR SHOWER PEAK TIME PREDICTION"
125 PRINT TAB(5);"----------------------------------"
130 PRINT
135 PRINT "WHAT IS THE YEAR (YYYY) ";
140 INPUT Y
145 PRINT
150 PRINT TAB(15);"METEOR SHOWER"
155 PRINT TAB(15);"-------------"
160 PRINT
165 PRINT " 1)  QUADRANTIDS";TAB(33);"1-4 JAN."
170 PRINT " 2)  APRIL LYRIDS";TAB(33);"20-23 APR."
175 PRINT " 3)  ETA AQUARIDS";TAB(33);"2-6 MAY"
180 PRINT " 4)  ARIETIDS";TAB(33);"1-15 JUNE"
185 PRINT " 5)  JUNE LYRIDS";TAB(33);"10-14 JUNE"
190 PRINT " 6)  DELTA AQUARIDS";TAB(33);"26-30 JULY"
195 PRINT " 7)  PERSEIDS";TAB(33);"10-14 AUG."
200 PRINT " 8)  ORIONIDS";TAB(33);"18-23 OCT."
205 PRINT " 9)  TAURIDS";TAB(33);"30 OCT.-10 NOV."
210 PRINT "10)  LEONIDS";TAB(33);"14-19 NOV."
215 PRINT "11)  GEMINIDS";TAB(33);"10-15 DEC."
220 PRINT "12)  URSIDS";TAB(33);"21-24 DEC."
225 PRINT
230 PRINT "TYPE IN THE NUMBER OF THE DESIRED SHOWER (1-12) ";
235 INPUT N
240 IF N<1 OR N>12 THEN 230
245 RESTORE
250 FOR I=1 TO N
255   READ S$,E,D$,A$,M,D
260 NEXT I
265 REM:  CALCULATE THE JULIAN DATE
270 J=365*(Y-1981)+INT((Y-1)/4)-INT((Y-1)/100)+INT((Y-1)/400)
275 J=J+INT(30.55*(M-1)-1.4)+2*INT((12-M)/10)+D
280 IF M<2 OR (Y/4)<>INT(Y/4) THEN 290
285 J=J+1
290 GOSUB 435
295 IF E1>E THEN 320
300   J=J+1
305   D=D+1
310   GOSUB 435
315   GOTO 295
320 IF E1<=E THEN 350
325   C2=E1
330   J=J-1
335   D=D-1
340   GOSUB 435
345 REM:  CALCULATE SHOWER PEAK TIME IN GMT
350 T=24*((E-E1)/(E2-E1))
355 H0=INT(T)
360 M1=INT(60*(T-H0)+.5)
365 G=100*H0+M1
370 IF D<=31 THEN 385
375 D=D-31
380 M=M+1
385 PRINT
390 PRINT "THE ";S$;" METEOR SHOWER WILL PEAK ON";M;"/";D;"/";Y-1900
395 PRINT "AT ";G;" GMT.  THIS SHOWER LASTS ";D$;" AND"
400 PRINT "THIS PREDICTION HAS AN ACCURACY OF ";A$
405 PRINT
410 PRINT"DO YOU WANT ANOTHER RUN (Y/N) ";
415 INPUT A$
420 IF A$="Y" THEN 115
425 GOTO 555
430 REM:  SUBROUTINE TO CALCULATE ECLIPTIC LONGITUDES
435 T1=(J-276)/36525
440 C0=0.016717-0.00004*T1
445 M2=(0.985609*(J-117.821)-0.3227*T1)/K
450 C2=M2+C0*SIN(M2)
455 M0=C2-C0*SIN(C2)
460 D0=(M2-M0)/(1-C0*COS(C2))
465 C2=C2+D0
470 IF ABS(D0)>0.0001 THEN 455
475 T=K*2*ATN(SQR((1+C0)/(1-C0))*SIN(C2/2)/COS(C2/2))
480 E1=T+1.7192*T1-77.396
485 IF E1<0 THEN E1=E1+360
490 RETURN
495 DATA QUADRANTIDS,282.83,10 HOURS,+/-15 MINS.,1,4
500 DATA APRIL LYRIDS,31.4,2 DAYS,+/- 12 HOURS,4,21
505 DATA ETA AQUARIDS,44.0,5 DAYS,+/- 12 HOURS,5,4
```

```
510 DATA ARIETIDS,75.0,8 DAYS,+/- 12 HOURS,6,5
515 DATA JUNE LYRIDS,84.0,2 DAYS,+/- 12 HOURS,6,14
520 DATA DELTA AQURIDS,125.0,2 DAYS,+/- 12 HOURS,7,26
525 DATA PERSEIDS,139.3,4 DAYS,+/- 75 MINS.,8,11
530 DATA ORIONIDS,207.0,2 DAYS,+/- 12 HOURS,10,20
535 DATA TAURIDS,220.0,20 DAYS,+/- 12 HOURS,10,31
540 DATA LEONIDS,234.7,3 HOURS,+/- 12 HOURS,11,16
545 DATA GEMINIDS,261.2,3 DAYS,+/- 12 HOURS,12,13
550 DATA URSIDS,270.0,12 HOURS,+/- 12 HOURS,12,21
555 END


        READY

        RUN

             METEOR SHOWER PEAK TIME PREDICTION
             ----------------------------------

        WHAT IS THE YEAR (YYYY) ? 1984

                        METEOR SHOWER
                        -------------

         1)  QUADRANTIDS                        1-4 JAN.
         2)  APRIL LYRIDS                       20-23 APR.
         3)  ETA AQUARIDS                       2-6 MAY
         4)  ARIETIDS                           1-15 JUNE
         5)  JUNE LYRIDS                        10-14 JUNE
         6)  DELTA AQUARIDS                     26-30 JULY
         7)  PERSEIDS                           10-14 AUG.
         8)  ORIONIDS                           18-23 OCT.
         9)  TAURIDS                            30 OCT.-10 NOV.
        10)  LEONIDS                            14-19 NOV.
        11)  GEMINIDS                           10-15 DEC.
        12)  URSIDS                             21-24 DEC.

        TYPE IN THE NUMBER OF THE DESIRED SHOWER (1-12) ? 7

        THE PERSEIDS METEOR SHOWER WILL PEAK ON 8 / 11 / 84
        AT  1939  GMT.  THIS SHOWER LASTS 4 DAYS AND
        THIS PREDICTION HAS AN ACCURACY OF +/- 75 MINS.

        DO YOU WANT ANOTHER RUN (Y/N) ? N

        READY
```

tic longitudes into the equation from above, we obtain:

$$T = 24 \cdot \frac{(E.L. - E.L.1)}{(E.L.2 - E.L.1)}$$

$$= 24 \cdot \frac{(234.70 - 234.28)}{(235.29 - 234.28)}$$

$$= 24 \cdot (0.416) = 9.98$$
$$= 0959 \text{ hours}$$

Checking back in history, we see that the peak indeed occurred on November 17, 1966, between 0900 and 1300 UTC.[11] Not bad for an estimate! Wait until the next Leonids peak in 1999; it will occur on November 17 at 1541 UTC. Check out the math yourself — and if I'm not around, remember that I told you so!

A few notes of caution are in order. As pointed out earlier, meteor shower peaks are based primarily on visual sightings. Radio peaks may vary slightly because the ionization for radio reflection may not necessarily cause a bright visual display. However, the radio and visual peaks are probably not much different. Another factor is the orbits of the showers themselves. Sometimes an orbit may be deflected, especially if it passes near one of the major planets such as Jupiter. If this happens, the shower may peak at a different date and/or time or even completely disappear! Also, some showers such as the Leonids are believed to be much younger, astronomically speaking, and hence are more concentrated. Therefore, they may be good only during certain peak years. The astronomy magazines are good sources of information to determine when these peak years will occur. In the interest of brevity, I have not mentioned minor showers or those that have been dormant in recent years perhaps because of a shift in orbit. They are still worth investigating if you have the time and inclination! New showers may be discovered but will take time to pinpoint exactly.

getting the best reflection

Another reason why different persons may quote different shower peaks is that they may have overlooked the geometry factor that must be observed.[4,6] In some cases there may be no optimum time for communications between two stations because a specific shower may not pass over the proper reflection point between the stations involved. However, when there is a usable meteor shower, there is an optimum time when the meteors will be in the proper location for the maximum signal reflection for the desired direction. W4LTU listed optimum times in his tables for scheduling between stations during meteor showers based on the direction of the path[6] This data is very important if you want to enhance your chance of a completed contact and not spend endless hours when the meteors are in the wrong position.

Some European VHF/UHFers have recently written computer programs to aid in selecting the best times to operate in any desired direction. They use the geometry specified by W4LTU in reference 4 with right ascension data for the meteor shower. Their program also adds data pertaining to effectiveness, thereby giving all times of the day with a quality factor of 0 to 100 percent for any desired path. Hopefully this program will soon be available here in the United States.

choosing the optimum shower

Another often overlooked fact is that each shower has its own specific characteristics. Time and space do not permit a long discussion here, but I will try to summarize some data I have collected to assist you in determining the best meteor shower to use for a specific frequency and path.

It is generally agreed that the opti-

table 2. Ecliptic longitude at 0000 UTC for selected days (from The American Ephemeris and Nautical Almanac for year of interest).*

date		1984*	1985*	1986*	1987*
Jan	3	281.79	282.54	282.26	282.04
Jan	4	282.81	283.55	283.27	283.06
Jan	5	283.83	284.57	284.29	284.08
Apr	21	31.08	30.81	30.54	30.34
Apr	22	32.06	31.78	31.52	31.32
Apr	23	33.03	32.76	32.49	32.29
May	4	43.72	43.45	43.18	42.99
May	5	44.69	44.42	44.15	43.95
May	6	45.66	45.39	45.12	44.92
June	5	74.53	74.26	73.99	73.80
June	6	75.48	75.21	74.95	74.75
June	7	76.44	76.17	75.91	75.71
June	14	83.13	82.86	82.60	82.41
June	15	84.08	83.82	83.56	83.36
June	16	85.04	84.77	84.51	84.32
July	27	124.15	123.88	123.63	123.43
July	28	125.11	124.84	124.58	124.38
July	29	126.06	125.80	125.54	125.34
Aug	11	138.51	138.25	137.99	137.79
Aug	12	139.47	139.21	138.95	138.75
Aug	13	140.43	140.17	139.91	139.71
Oct	20	206.78	206.52	206.52	206.04
Oct	21	207.78	207.51	207.24	207.03
Nov	2	219.76	219.49	219.22	219.00
Nov	3	220.76	220.49	220.22	220.00
Nov	4	221.76	221.49	221.22	221.00
Nov	16	233.82	233.55	233.28	233.06
Nov	17	234.83	234.56	234.28	234.07
Nov	18	235.84	235.57	235.29	235.08
Dec	12	260.15	259.88	259.61	259.39
Dec	13	261.17	260.90	260.62	260.41
Dec	14	262.19	261.92	261.64	261.42
Dec	21	269.31	269.04	268.76	268.55
Dec	22	270.33	270.06	269.78	269.56
Dec	23	271.35	271.08	270.80	270.58

*Each table repeats itself every 4 years. Hence the table for 1984 will be usable in 1988, 1992, and so on. See text for further information.

mum path length for meteor scatter communications is between 700-1000 miles (1126-1609 km), but this is based mainly on an average meteor's ionizing at 100 km altitude. The slower velocity showers (11-40 km/sec., see velocity column on **table 1**) such as the Geminids can ionize as low as 60-80 km and hence are good for shorter distances, while the faster showers (40-72 km/sec.) such as the Perseids may ionize at as high an altitude as 150 km. Therefore, the faster showers will generally yield the best DX.

Likewise, the fast showers seem to be the best ones for 220 MHz and above. The duration and signal strength of the burst drops off rapidly with decreasing wavelength and therefore is considerably shorter and weaker on 220 MHz than on 2 meters. Even on a fast shower, very few long bursts will be detected on 70 cm. The ideal shower for the most difficult paths and higher frequencies should have a good rate of meteors per hour because several bursts may be required to complete the exchange of information required for a QSO. Therefore, fast showers with few meteors, such as the Leonids of late, are not good prospects except during peak years in their cycle. The Orionids and Eta Aquarids meteor showers are believed to be associated with Halley's comet which is due to rendezvous with the sun in 1986. This could mean a big increase in the rate of these showers over the next few years. Stay tuned in!

In short, I feel the Perseids is the ideal all-purpose shower, especially for 220 MHz and above. It has a long duration, high speed, lots of large particles, and is evenly distributed from year to year. Is it any wonder why all reported 70 cm meteor scatter completed contacts have taken place during this shower? The Geminids, even though slow in speed, are reliable year after year for paths up to 1000 miles (1609 km), with lots of meteors and good performance even on 220 MHz. The Quadrantids are great if you are lucky enough to catch the narrow peak. The Delta Aquarids can also be good, but it has been pointed out in recent years in astronomy magazines that this shower may be comprised of up to six or seven other showers, hence explaining its erratic nature from year to year.

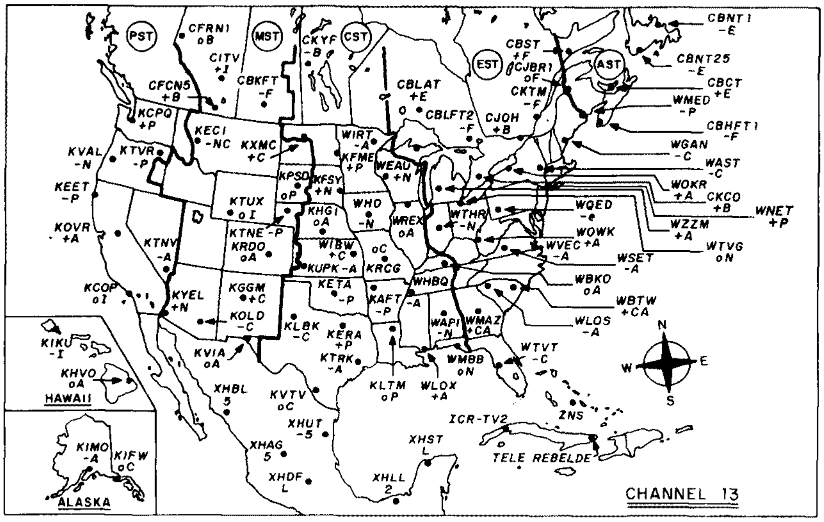


fig. 2. Map showing TV channel 13 locations and offset can be used to determine meteor scatter openings. (Reprinted with permission from the Worldwide TV FM DX Association, Buffalo, New York.)

optimizing your equipment

It is possible to make 6- and 2-meter meteor scatter contacts during meteor showers with power as low as 10 watts and a 10-12 dBd gain antenna. However, these contacts would not be the long-haul DX (over 1200 miles or 1931 km) and would take some luck. For reliable meteor scatter work on 6 and 2 meters, it usually requires greater than 100 watts and an antenna gain of 12-15 dBd. A good noise figure would be desirable to take advantage of the weaker and shorter bursts, but even a 5-10 dB noise figure will work because bursts on these frequencies are usually well above the noise.

QSOs on 220 MHz and 70 cm via meteor scatter are another story. Signal levels are definitely weaker, especially on the longer haul. The bursts are also much shorter. A "blue whizzer" on 2 meters may be 30 to 90 seconds long, but the same kind of burst on 220 MHz may last only 15 to 30 seconds. On 70 cm, pings are typical, and if you're lucky enough to get a burst, it will probably only be a few seconds long since 5 to 10 second bursts are extremely rare. Hence, it helps to have a good receiver noise figure (2 dB or better) and more power, at least 100 but 500 watts or more if possible. Likewise, higher antenna gain is desired.

A factor often overlooked on meteor scatter communications is the geometry involved. It has been pointed out by W4LTU that for ideal conditions, the reflection point from the meteor trail will be optimum approximately 5 to 9 degrees on either side of the direct path depending on path length.[4] Fortunately, as the path lengthens, the angle narrows. Indeed, it has been further suggested that you can optimize the path by aiming your antenna to the proper offset from the direct path as pointed out in **table 1** of reference 6. Antenna height is also important, because nulls will appear in the pattern based on height in wavelengths. Herein lies the rub. More antenna gain will only aggravate the situation because higher gain implies narrower beamwidth.[12]

On 220 MHz the typical high gain Yagis in present-day use have a beamwidth of 26 to 30 degrees. If these antennas are stacked two high and/or two wide, the beamwidth will typically be 13-15 degrees, meaning that signals at the reflection point on either side of the path will be at about 3 dB down. On 70 cm the problems are usually worse; typical high performance stations use four Yagis, each with 20 to 24 degree beamwidth. An array of four Yagis arranged in a square would be 10 to 12 degrees wide, at least 3 dB down on the band where we are already fighting feedline loss and very weak, short duration bursts.

There is a partial solution to this dilemma. If possible, stack your antennas vertically so that the horizontal beamwidth will be the same as a single antenna. The beamwidth may drop to 4 to 6 degrees in the vertical plane, but it doesn't really matter at distances greater than 750 miles (1207 km).[4]

monitoring a meteor shower

There is an easy and effective way to monitor meteor showers. I have been listening to VHF TV station channels 12 or 13 video carriers for many years with a simple homebrew converter and Yagi antenna.[13] The normally specified TV video carrier frequency is 1.25 MHz above the bottom end of the allocated band. What isn't always known is that on the VHF channels (2 through 13), there are three different video carrier frequencies for each TV channel: one on the listed frequency, called the zero offset; one 10 kHz below, named the minus offset; and one 10 kHz above, labeled the plus offset. The accuracy of these carriers is ± 1 kHz and they are assigned by the FCC so as not to cause video beating between similar channels in any wide geographical area. I have listed the standard carrier frequencies in **table 3**. (The actual assigned frequency for each TV station is listed in *Part 73, Subpart E-Television Broadcast Stations of the FCC Rules and Regulations.*) Knowing the assigned frequency for a specific station will aid in positive identification. These frequencies, along with offsets, effective radiated power, and such are also listed in the *WTFDA TV Station Guide*,[14] with individual maps for each TV channel to aid the user to locate the stations on each channel (see **fig. 2**).

Several things should be considered when selecting a TV channel to monitor. A TV channel 2 monitor (55.25 MHz) may be very active and useful for 6 meters, but gives little insight into what's really happening up on 2 meters or 220 MHz. It is best to use one of the

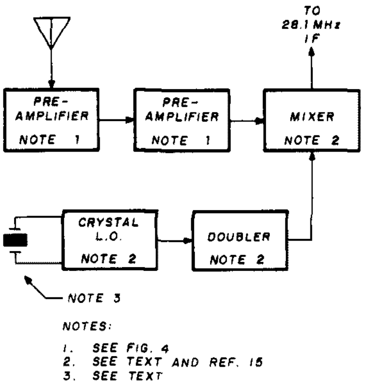


fig. 3. Block diagram of a typical receive type down-converter for monitoring TV video carriers.

table 3. Selected TV video carrier frequencies for U.S.A., Canada, and most North American TV stations.

TV channel No.	frequency (MHz)*
2	55.250
3	61.250
4	67.250
5	77.250
6	83.250
7	175.250
8	181.250
9	187.250
10	193.250
11	199.250
12	205.250
13	211.250

*This is the zero-offset frequency. Channels may be on this frequency, ± 10 kHz of the same based on FCC frequency assignments (see text).

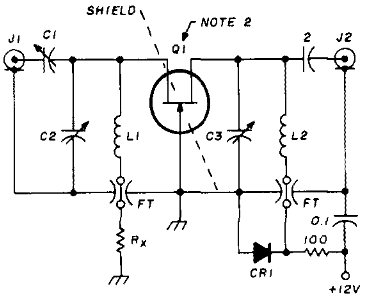
higher VHF channels since they are widespread; this gives plenty of opportunity to hear meteor bursts in different areas. The higher frequency channels also give a better indication when the meteors are really hot for DX, and when 220 MHz is possible. Channels 12 and 13 are ideal in this regard. Select a channel that isn't used locally (within 100 miles or 161 km), because strong channels could overload your receiver and may produce extraneous 15,750 kHz video birdies.

My converter is typical of those used on 2 meters or 220 MHz for weak signal work and is patterned according to the circuitry shown in my March, 1984 *ham radio* article.[15] A block diagram of this converter is shown in **fig. 3**. I chose channel 13 (video carrier on 211.250 MHz from **table 3**) and use a local oscillator frequency of 183.150 MHz derived with the actual crystal oscillator operating at 91.575 MHz. My IF is 28.1 MHz for reasons specified in reference 15. For a preamplifier I used two U310 JFET stages as shown in **fig. 4** which will work from 200-225 MHz and over a wider frequency range if the inductors are scaled accordingly. This configuration is more than adequate to hear the weak TV video carriers out to 300 miles even on poor days with no propagation enhancements. Remember, these stations are radiating high power and usually at high elevations.

The choice of receiving antenna is very important. For best operation you should have a sharp and clean antenna pattern with low side lobes so that you can null out a loud station as well as be able to distinguish which direction the signals are coming from. I prefer the NBS 2.2 wavelength Yagi and have shown a suitable design in reference 16. I have recently changed to a T-match and would recommend same if you are making your own antenna.

A few final comments about monitoring TV video carriers: short meteor bursts on channel 12 or 13 will probably indicate that 2-meter scatter is good; 5 to 10 second bursts probably mean good DX as well as opportunities on 220 MHz and above. I have noticed that the burst starts first on the higher frequency. Hence, if you hear a good burst start in the desired direction, the lower frequency path will open shortly afterward. Conversely, a burst on 200 MHz may occur too late for you to catch a 70 cm path. Pings or rapid Doppler means that there are underdense bursts which aren't going to be of much help, but they do indicate that there are meteors present. Use of this type of monitor is not only helpful to see when, where, and if a meteor shower is in progress, *but can also be used to catch other VHF/UHF openings such as sporadic E, auroral or tropo.*

operating procedures



Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms.
k = 1,000 M = 1,000,000

item	description
C1,C2,C3	1-10 pF air variable capacitors
CR1	1N4001 or equivalent "idiot" diode
FT	0.001 μF feedthrough capacitor
J1,J2	BNC connector
L1,L2	3 turns No. 18 1/4" ID, 1/2" long
Q1	U310 JFET (do not substitute J310) (see Note 2)
R_x	set for I_D 8-12 mA, 150 ohms typical

Note 1. Preamplifier can be built on P.C. board mounted to cover of a Pomona 2417, Bud CU123 or equivalent cast box per construction techniques in reference 15.
Note 2. Mount U310 upside down in a 0.191" (4.85 mm) diameter hole (made with a No. 11 drill). Quickly solder the transistor can to the P.C. board.

fig. 4. Low noise JFET preamplifier (200-225 MHz) for video carrier monitor has a 11-12 dB gain, a noise figure of 2.5 dB and a 8-9 MHz bandwidth.

In the United States, meteor scatter schedules are usually run for one hour with the station farthest south and/or west transmitting the first and third 15 seconds of each minute. CW used to be the prime mode, but now SSB is used almost exclusively. CW is best on weak signals (especially for long DX and 70 cm) and is easy to tune in, but has a low data rate. SSB is definitely faster — especially if the burst is long enough to complete the exchange — but is more difficult to tune in especially when signals are weak. Call signs must be exchanged by both stations although not necessarily in a single burst. Reports and "rogers" received by both parties complete the QSO. If a burst is long or falls at the end of a transmission, break-in procedures may be used to complete the contact on a single burst.

Until recently, Europeans had only two bands to use for meteor scatter, 2 meters, and 70 cm. In Europe, high speed (75-125 WPM!) CW is still in wide use. In contrast to the United States, where the contact takes place in real time, the Europeans usually run long (1 to 5 minutes) transmitting sequences and then record on tape. This

method is less critical on timing (using 15 second sequencing can cause a 33 percent loss of information if one station is off 5 seconds in timing!). At the end of a receiving sequence, the tape is quickly replayed at reduced speed to pick off any information received. This procedure does favor short bursts because entire call sets and reports can be sent in seconds or less and replayed until the information is deciphered. SSB is now gaining popularity in Europe.

In some sense, the high-speed CW used in Europe is similar to the procedures being proposed by those interested in packet or computer-oriented communications.[17] It's no secret that many of us still feel that communications must still be heard and deciphered by the operator rather than just appear as a letter or message on a video screen. However, such is progress! Reviewing the material in reference 17 should be very interesting to those with computer experience.

summary

I can't believe this article grew so long! Hopefully the material presented will be interesting and helpful. Let me point out that I am not advocating the use of only certain meteor showers. I presented this material to help you increase your success record — especially on the longer DX, 220 MHz and above. If the shower isn't materializing, why waste the kilowatts? However, using the material presented here, you may be able to discover other shower opportunities or input data to refine the peaks of existing showers.

Various methods have been presented to predict more accurately when the better meteor showers will peak. Remember, you can't use a short duration shower if it peaks when the radiant is not above your horizon! In addition to the information in **table 1**, refer back to W4LTU's table[6] for best times for the path direction desired. Optimize your equipment especially if it is not up to the standards recommended; build up a monitor as described and you'll be amazed how helpful it will be, not only for meteor scatter but for other propagation modes as well.

Finally, let me thank W4WD and AD1C for their help in determining the shower peaks by computer. Also, many thanks to Chip Brown, KR1P, for the material he helped me obtain on this subject as well as his helpful suggestions; to WTFDA for use of their map; and the people at *Sky and Telescope Magazine* for their help. Good luck on your next meteor scatter schedule and let me know if you can improve the data base in this article.

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VHF/UHF coming events

June 5:	*1200 UTC, peak of Arietids daytime meteor shower*
June 9-10:	*ARRL VHF QSO Party*
June 14:	*2200 UTC, peak of June Lyrids meteor shower*
June 15-17:	*SMIRK (6 meters) Contest*
June 21:	*Mean date of 2-month annual peak of sporadic-E propagation*

ham radio

the peaked lowpass: a look at the ultraspherical filter

Use a pocket calculator for filter design, analysis, experimentation

Much of the fundamental work done in filter design and analysis over the past twenty years has been the result of detailed computer studies done by researchers in industry and education. Now, with personal computers, even home hobbyists can perform the complex calculations necessary for thorough examination of existing filter types and understanding of new types, exceeding the current standards, as they become available.

The most obvious opportunity for Amateur exploration in this field is in design, using published equations for the generation of tables. For example, using modern network theory,[1] a computer program can easily generate the table for Chebyshev filters of any arbitrary ripple value.[2] This allows one to design using available standard components.[3]

An equally useful computer application is filter analysis. An analysis program allows the user to enter component values and examine the response of that filter. Both gain (transducer gain) and phase response are easily generated with a simple ladder analysis program. More refined programs provide details regarding the filter response to a nonsinusoidal input such as an impulse or step; each of these is of interest to the circuit theorist.[4]

A well-designed filter analysis program does more than confirm the traditional — and therefore expected — results. It allows practical details, such as the effects of loss in the elements (finite Q inductors and capacitors) and parasitic reactances, to be enumerated and evaluated. One can also use a filter analysis program to study what happens when certain component values are changed.

It isn't even necessary to use a personal computer; for this study, a modest but powerful handheld calculator (an HP-41CV) was employed, using a previously published ladder analysis program.[5,6] The program was applied to the ultraspherical filter, a polynomial filter whose utility in certain special applications outweighs its apparent lack of popularity. (The fundamental work on this type of filter was done by Johnson and Johnson;[7] their work will be extended — and some practical details added — in this article.)

lowpass filters

To begin, let's examine a traditional lowpass filter. The filter chosen, a 5-pole Chebyshev with a 2-dB passband ripple, has been designed for a ripple *cutoff* frequency of 5.258 MHz. Its circuit appears in **fig. 1** with its frequency response shown in **fig. 2.** The usual Chebyshev characteristics are evident. There are five half-cycles of ripple in the passband. The response is down by 2 dB, the ripple value, at the 5.258 MHz ripple cutoff frequency.

This filter also exhibits an interesting but rarely appreciated property of the Chebyshev: with a 5.258-MHz ripple cutoff, the filter has its final peak at 5 MHz. Further examination reveals that the relationship between the positions of the response peaks are related to the cutoff by constants that are *not* dependent on the ripple. That is, all doubly terminated 5-pole Chebyshev filters will have their final peak at a frequency that is below the ripple cutoff by a factor 0.9509 (= 5.00/5.258) for all ripple values. Further analysis provides the relationships for other filter orders and for the other peaks.

Having examined a classic filter, let's now consider the effect of changing, or perturbing, some of the

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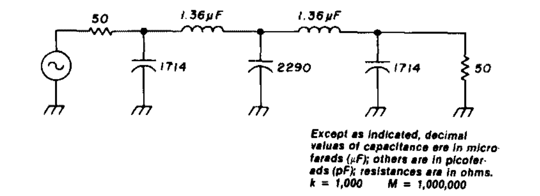


fig. 1. Schematic of a 2 dB ripple Chebyshev lowpass filter with ripple cutoff at 5.258 MHz.

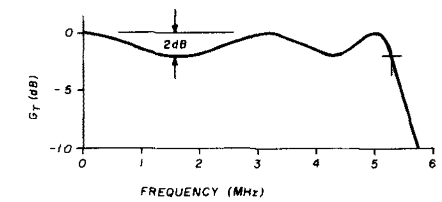


fig. 2. Frequency response of a Chebyshev lowpass filter with 2 dB passband ripple.

component values. First, we'll increase the inductor values from those shown in **fig. 1**. We'll keep the filter symmetrical with equal inductors, but increase the value from 1.36 to 3 microhenries. With the larger inductors, we would expect the cutoff frequency to decrease; this is confirmed in the frequency response (**fig. 3**). Of greater interest, however, is a change in filter shape. The filter still has ripples in the passband, but they are no longer of equal value. The first (close to zero frequency) dip in the response is not as deep as the second one, closer to the cutoff. A further increase in inductor values will further emphasize this effect.

Next we examine the modified filter with the goal of moving the response peak back to 5 MHz. This can be realized in at least two ways. All component values can be reduced equally to produce a filter with an identical shape with the peak at 5 MHz. We could, alternatively, reduce only the center capacitor until the final peak is at 5 MHz. This occurs with the center capacitor reduced from 2290 to 812 pF, yielding the response shown in **fig. 4**. This result is especially interesting — the shape is an even more extreme departure from the original Chebyshev, with a drastic difference in the dips in the curve — the ripples. Indeed, the smaller ripple, shown in **fig. 4**, is labeled in order that it not be missed!

This filter is so extreme that we might well ask whether it's worth considering for any practical application. The ripple has become large enough that the filter has taken on a bandpass-like filter characteristic. This wouldn't be very useful for an application in which a true lowpass shape was needed. However, this is *not* the way lowpass filters are used in most applications; in Amateur Radio, the most common application is in harmonic filtering, in which we wish to pass one frequency or narrow band while attenuating the harmonics of that frequency as much as possible. The filter with the response shown in **fig. 4** will do just that. The harmonic attenuation is greater at 10 MHz than the original Chebyshev prototype. We'll term this circuit a "peaked lowpass filter," or PLPF.

Returning to the original Chebyshev prototype of **fig. 1**, we can now consider an opposite pertubation,

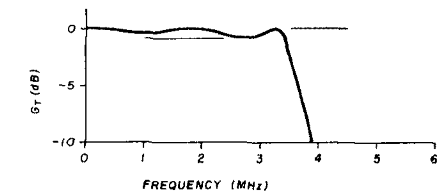


fig. 3. Response of an ultraspherical filter. Note that the ripples are not equal.

a reduction of inductor values. Allow the inductors to decrease from 1.36 to 1 microhenry; this will increase the peak frequency above the original 5 MHz. This increase is compensated for by increasing the center capacitor from 2290 to 4000 pF. The resulting response and circuit are shown in **fig. 5**. Again, the response has five half-cycles of ripple oscillation. The ripples are unequal, but the large dip in the passband is now the one close to zero frequency. This filter might be termed a "double peaked lowpass." This particular circuit, if properly scaled, would produce a lowpass response with good harmonic attenuation, and with a flat characteristic with only 0.15 dB ripple over the 3.1 to 4 MHz range.

insertion loss

The frequency response curves presented so far have assumed lossless components. (This is the common approach in analyzing lowpass filters.) While this can be a mistake, it is easily remedied with the computer or calculator. A more detailed examination will show that the main loss occurs in the inductors. Typically, the inductors have Q_u (unloaded Q) values of 250 or less (toroids at HF). In contrast, mica capacitors often show a Q_u value in excess of 700 at HF.

If we assume a Q_u that is constant with frequency, we find that the greatest loss occurs at or near the lowpass filter cutoff. (Effects are minimal at lower frequencies.) This loss can be significant in a high-ripple Chebyshev. The effects are also significant in PLPF,

especially as the peak becomes sharper. Design then becomes a tradeoff between acceptable insertion loss and the desired selectivity characteristics. This tradeoff between loss and selectivity is a central theme in all filter design, and is especially significant with band-pass filters with a narrow fractional bandwidth.

the double pi-network: a design method for the PLPF

The circuit we have discussed is a familiar one if we consider only superficial details. It could be a lowpass designed from tables for one sort of polynomial or another. It is, however, a double pi-network, the cascade of two familiar impedance matching networks. The matching circuit is designed for a single specific frequency; our analysis of the PLPF has also focused on a single peak frequency. This similarity suggests a method for design.

First, we'll analyze the impedance characteristics of a pi-network. Our analysis will then form the basis for derivation of some special design equations for the PLPF. This method may be used for higher order versions of the PLPF, either with numerical methods on our computer or in closed mathematical form.

Analysis of the pi-network begins with the circuit of **fig. 6A**, an end resistance, R_o, paralleled with an end capacitor, C_e. The admittance of this combination is written directly, $Y_a = G + j\omega C_e$ where $G = 1/R_o$. This is converted to an impedance by taking the reciprocal of Y.

The impedance looking into plane "a" may now be combined with a series inductor, shown in **fig. 6B**. The resulting impedance is $Z_b = Z_a + j\omega L$. Keep in mind that both Z values are complex, with both real and imaginary parts. The admittance looking into plane "b" is obtained again by complex inversion, $Y_b = 1/Z_b$. This admittance consists of two parts, a real conductance, G_b, and an imaginary susceptance, jB_b. The imaginary term will be negative, indicating that the admittance is inductive. This may be cancelled, or tuned with a shunt capacitance, C_t, producing the traditional pi-network of **fig. 6C**. The capacitance required for tuning is:

$$C_t = \frac{-1}{\omega} Im(Y_b) \qquad \textbf{(1)}$$

where $\omega = 2\pi f$ and f is in Hz. L and C values are in Henrys and Farads.

Thus we have formed a traditional pi-network. The input impedance looking in will be real (resistance only), with a value of $R_{in} = 1/G_b$. This circuit would most likely be mismatched if driven by a source resistance R_o. However, we can eliminate the mismatch, and the loss that would result from it, by transforming an R_o source to "look like" a source resistance of R_{in}. This is easily realized with another pi-network identical to the one we have just "de-

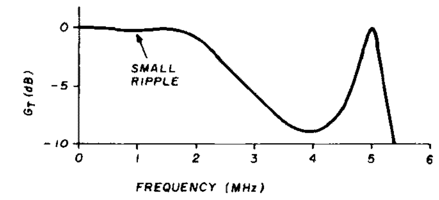


fig. 4. Response of extreme version of a PLPF.

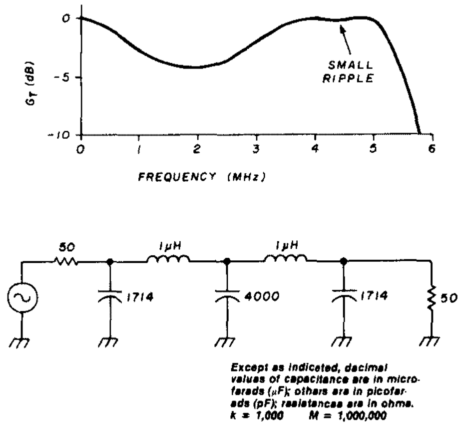


fig. 5. Double peaked lowpass, ultraspherical filter. Note the small ripple between 4 and 5 MHz.

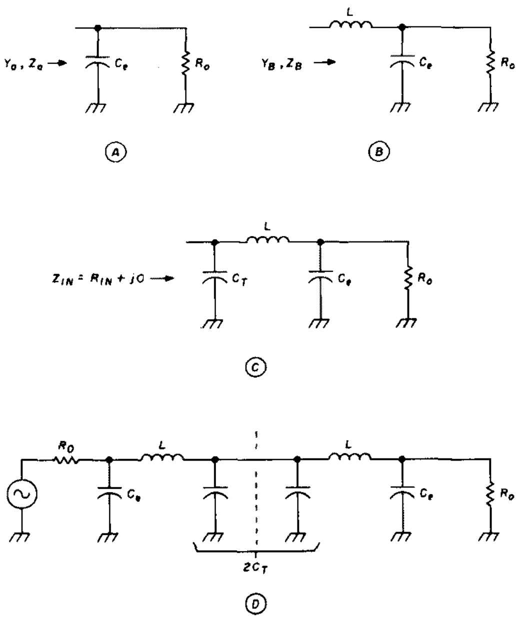


fig. 6. Circuits showing the analysis of a pi-network and its evolution into a double pi-network.

table 1. Normalized component values for 5th order ultraspherical lowpass filters. $Q_u = 200$ for inductors, capacitors are lossless. Also shown is the peak ripple in the passband, the insertion loss and the responses at the 2nd and 3rd harmonics.

g_1	g_2	g_3	ripple dB	IL, dB	$-G_t(2f_p)$ dB	$-G_t(3f_p)$ dB
1	1.0	2.0000	0.37	0.09	25.1	45.7
1	1.2	1.8919	0.16	0.10	29.0	48.8
1	1.4	1.6981	0.21	0.12	31.4	50.9
1	1.6	1.5068	0.47	0.14	33.2	52.3
1	1.8	1.3402	0.84	0.16	34.5	53.5
1	2.0	1.2000	1.29	0.17	35.6	54.5
1	2.5	0.9412	2.56	0.22	37.7	56.4
1	3.0	0.7692	3.83	0.26	39.4	57.9
1	3.5	0.6486	5.03	0.30	40.7	59.2
1	4.0	0.5600	6.13	0.34	41.9	60.3
1	5.0	0.4390	8.04	0.42	43.8	62.1
2	1.0	3.0000	1.15	0.22	42.2	61.8
2	1.2	2.3529	1.78	0.26	43.6	63.0
2	1.4	1.9231	2.82	0.30	44.7	64.0
2	1.6	1.6216	3.84	0.34	45.7	64.9
2	1.8	1.4000	4.81	0.38	46.5	65.8
2	2.0	1.2308	5.71	0.42	47.3	66.5
2	2.5	0.9438	7.70	0.53	49.1	68.1
2	3.0	0.7647	9.35	0.63	50.5	69.5
2	3.5	0.6425	10.76	0.73	51.7	70.7
2	4.0	0.5538	11.99	0.83	52.8	71.8
2	5.0	0.4340	14.02	1.02	54.6	73.5
3	1.6	1.5294	8.99	0.67	52.3	71.5
3	1.8	1.3274	10.11	0.75	53.1	72.3
3	2.0	1.1724	11.10	0.83	53.9	73.1
3	2.5	0.9072	13.20	1.02	55.7	74.8
3	3.0	0.7397	14.89	1.21	57.1	76.2
3	3.5	0.6244	16.31	1.40	58.4	77.4
3	4.0	0.5401	17.54	1.58	59.4	78.5
3	5.0	0.4253	19.57	1.94	61.3	80.3
4	1.6	1.4628	13.51	1.11	56.9	76.1
4	1.8	1.2764	14.63	1.24	57.8	77.0
4	2.0	1.1321	15.62	1.36	58.6	77.8
4	3.0	0.7231	19.39	1.97	61.9	81.0
4	4.0	0.5311	22.01	2.54	64.2	83.3
4	5.0	0.4197	24.02	3.08	66.1	85.2
5	1.6	1.4197	17.24	1.64	60.6	79.8
5	2.0	1.1059	19.33	2.01	62.3	81.5
5	3.0	0.7122	23.05	2.86	65.6	84.7
5	4.0	0.5252	25.65	3.64	68.0	87.1
5	5.0	0.4160	27.65	4.35	69.9	89.0

signed," but reversed. The final circuit is shown in **fig. 6D**.

We have done all our design and analysis at one frequency, one where we desire a response with no loss (assuming ideal components). This is not especially illuminating as far as filter properties are concerned. Still, it has provided a simple method for choosing the center capacitor in a filter so as to ensure minimum attenuation at one particular frequency.

A direct extension of the previous analysis will lead to design equations for the 5th order PLPF. If the end capacitor chosen has reactance X_e at the peak frequency and the inductor has reactance X_L, with the end resistance R_o, the total center capacitance is:

$$C_t = \frac{1}{\pi f}\left[\frac{X_L - \dfrac{R_o^2}{X_e}(1 - X_L/X_e)}{R_o^2(1 - X_L/X_e)^2 + X_L^2}\right] \quad \textbf{(2)}$$

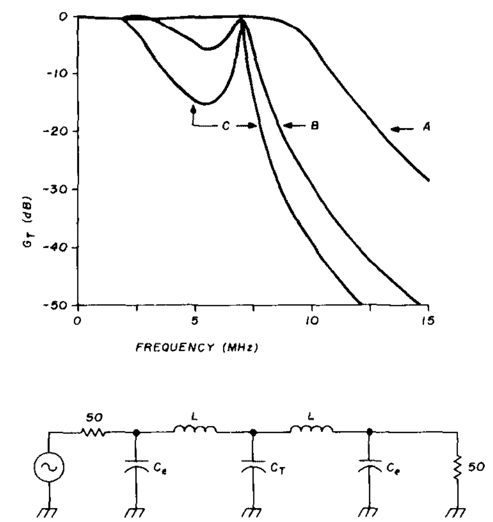
where f is the peak frequency in Hz

We have already mentioned the importance of considering insertion loss. Further manipulation of the equations (with the assumptions that capacitors are lossless, but that the inductors have a known unloaded Q), shows that the insertion loss is:

$$IL\ (dB) = 20 \log\left[1 + X_L\frac{X_e^2 + R_o^2}{X_e^2 R_o Q}\right] \quad \textbf{(3)}$$

These equations may be used for the design of a practical PLPF if some restrictions are noted. The end capacitor should have a reactance $X_e \leq R_o$, while the inductors should have a reactance $X_L \geq R_o$. Both are evaluated at the peak frequency.

Filter design usually begins with standard-value end capacitors and inductors already on hand. The value of the tuning capacitor and level of insertion loss are calculated; if the loss is acceptable, construction can proceed. If an appropriate program is used, analysis will provide the expected levels of harmonic attenuation.



CASE	C_e	L	C_T
A	455	1.14μH	903
B	1000	2μH	647
C	1500	3μH	385

Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

fig. 7. Three variations of the PLPF produce dissimilar responses.

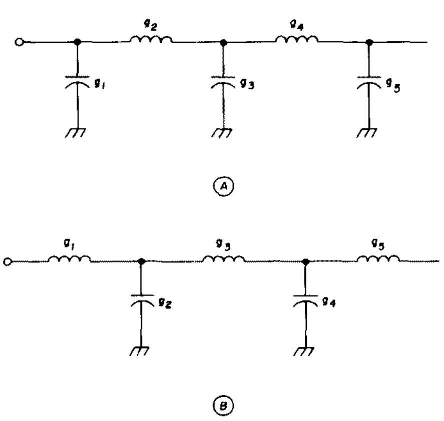


fig. 8. Normalized peaked lowpass filters (PLPF) to be used with *table 1*.

The nature of the PLPFs possible is illustrated in **fig. 7**. Three filters have been designed for a peak at 7 MHz. One, shown as curve A, is the popular half-wave filter.[9] While its peak is not very sharp, this is an easy filter to build. The insertion loss for inductors with $Q_u = 250$ is low, but the harmonic attenuation is poor at best. This filter may be designed with methods outlined, with $X_e = X_L = R_o$. Curves B and C of **fig. 7** are much more dramatic, showing significantly improved harmonic attenuation. The insertion loss is higher for these filters, however, but this is the price of increased selectivity.

normalized tables for the PLPF

Most of the more familiar lowpass filters are designed from tables. Extensive data is available for the Chebyshev and Butterworth polynomial filters, as well as many more.[10] Manipulation of the equations above allows us to formulate tables for the PLPF, presented in **table 1** for only the 5-element filter. The g_k values shown are the values of capacitors or inductors for a filter with a peak frequency of 0.1592 Hz (one radian) and 1 ohm terminations at both ends. The possible circuits are shown in **fig. 8**. The form with capacitors at the ends is preferred. The first and second elements, g_1 and g_2, were chosen arbitrarily. Then the required tuning element, g_3, was calculated. **Eq. 2** was placed in normalized form for this evaluation,

$$g_3 = \frac{2(g_1^2 g_2 + g_2 - g_1)}{(1 - g_1 g_2)^2 + g_2^2} \qquad \textbf{(4)}$$

Table 1 contains the normalized component values for several PLPF circuits. Also shown are the values of the largest ripple in the "passband," the insertion loss is based on an inductor Q of 200, and the attenuation values at the 2nd and 3rd harmonics. It's easy to use the table for designing a PLPF; simply study the insertion loss and harmonic attenuation columns to find a circuit that will meet the requirements for the application at hand. Then design the circuit using,

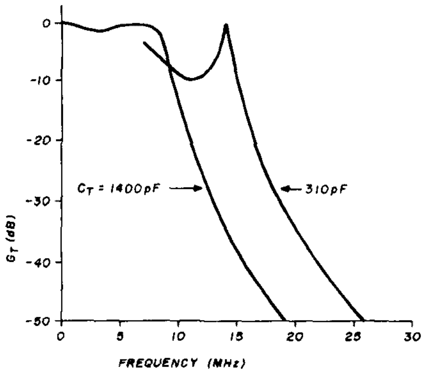
$$L = \frac{g_2 R_o}{\omega_p} \qquad \textbf{(5)}$$

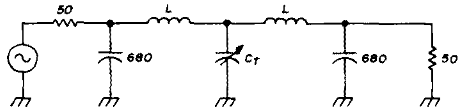
and

$$C_k = \frac{g_k}{R_o \omega_p} \qquad \textbf{(6)}$$

where $\omega_p = 2\pi f_p$ with f_p being peak frequency in Hz. **Equation 6** provides both the end capacitors from g_1 and the tuning capacitor with g_3.

Any filter designed directly from the table would probably not be the most practical one. A more realistic design method would use the table merely as a guide: a preliminary filter would be designed according to the table; end capacitors and inductors — with





L = 1µH
Q_U = 250

Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

f_{PEAK}	C_T	IL	G_T (AT $2f_{PEAK}$)
7 MHz	1400	0.1dB	-34.1dB
10 MHz	698	0.24dB	-46.0dB
14 MHz	310	0.59dB	-52.9dB
21 MHz	125	1.75dB	-62.8dB
28 MHz	68	3.62dB	-70.0dB

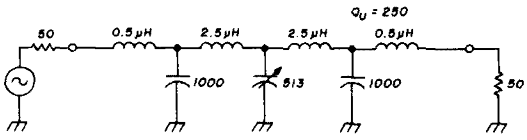
fig. 9. A PLPF that may be tuned over a 4:1 frequency range with the change of but one element, the center capacitor. (See text for discussion.)

values close to those calculated — would then be selected from the junk box. **Equation 2** would be used to determine the final tuning capacitor value.

There is one practical complication associated with the PLPF, especially in the more extreme examples, in which g_1 and g_2 are large in comparison to unity. The peak frequency will change significantly with the tuning capacitor. Hence, a trimmer is usually required. This is a small price to pay for the improved harmonic attenuation that can result.

variations on the PLPF

Many things can be done with the PLPF. One problem mentioned in the previous section was the tuning sensitivity of the center capacitor. An adage used by engineers applies to this situation: "If you can't fix it, feature it." We find that a peaked lowpass filter may be tuned over a wide frequency range merely by changing the center capacitor. This is illustrated with the filter and response curves shown in **fig. 9**. A single filter is tuned over the range of 7 to 28 MHz with one variable element. The ripple depth and the insertion loss both grow as the filter is tuned to higher frequencies, limiting its practical application.



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

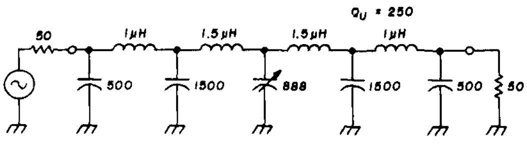


fig. 10. Ultraspherical filters of 7th and 9th order.

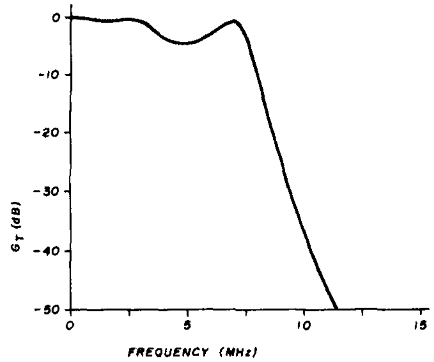


fig. 11. Frequency response of the 9th order lowpass filter shown in *fig. 10*. Note the combination of unequal passband ripple and good harmonic attenuation.

The ultraspherical filter is not limited to circuits with only five elements. **Fig. 10** shows two circuits peaked at 7 MHz using 7 and 9 elements. Methods such as those used in analysis of the pi-network were applied to determine the value of the tuning capacitors in these circuits. The 7-element filter has an insertion loss of 0.38 dB and 2nd harmonic attenuation of 52.5 dB. The 9th order filter has an even lower IL of 0.30 dB and 2nd harmonic attenuation of 71.4 dB. The reduction in IL results from the improved impedance matching aided by the additional end elements. The matching is effective at the peak frequency, but continues to contribute to the filtering at harmonics. **Fig. 11** shows the response of the 9-element filter of **fig. 10**.

An interesting traditional filter is the elliptic or Cauer-Chebyshev.[11] A lowpass of this type may have its inductors replaced by parallel resonant traps. We can perform the same modification on the PLPF. An ex-

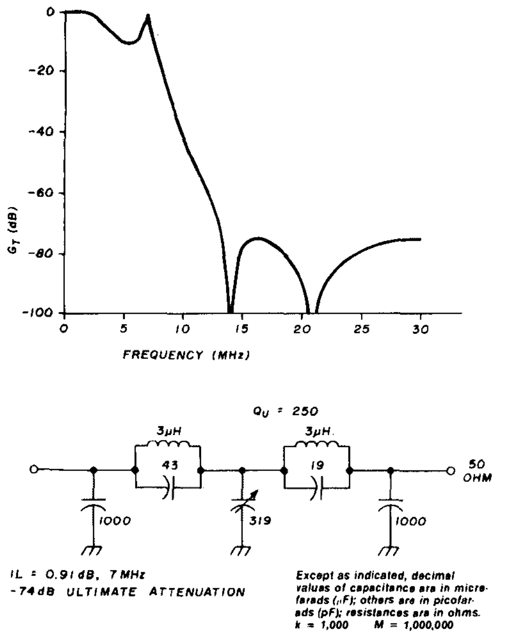


fig. 12. Application of elliptic filter methods to a 5th order PLPF.

ample is presented in **fig. 12**, peaked at 7 MHz. The traps are tuned to 14 and 21 MHz. The result is a filter similar to the ultraspherical in the passband with unequal ripples, but with deep notches at the harmonics. The main deficiency of this circuit is a limited ultimate attenuation at VHF of −74 dB. This limitation is typical of traditional elliptic circuits.

bandpass filters based upon the PLPF

A characteristic found with the peaked lowpass circuit was a bandpass-like response for the extreme cases. This characteristic can be used to advantage, especially when a filter must also have extremely good attenuation in the high frequency stopband.

Fig. 13 shows a filter that was designed as a preselector to precede an 80-meter receiver.[12] The receiver used a 9 MHz IF and a 5 MHz LO, so the image was at 14 MHz. High image rejection was required, and some "close-in" selectivity was required in the 3.5 to 4 MHz region. A 5th order PLPF was used for the basic filter. This was cascaded with a 7th order highpass circuit to eliminate the low frequency response. **Fig. 13** gives the overall response as well as those of the lowpass and highpass sections alone. Note the bump in the bandpass response at the 3 MHz cutoff of the highpass. This was not a problem in practice. The overall filter yielded nearly 100 dB image rejection and was tunable over the 80-meter band with a single section variable capacitor.

The more interesting bandpass filters are those with a multiplicity of resonators, or tuned circuits. Hence, two of the previous PLPF circuits were cascaded (on the computer), producing the response and circuit shown in **fig. 14.** The response shows a double peaked response like that of an overcoupled double tuned circuit.[13] The 0.01 μF coupling capacitor between PLPF sections was increased to 0.015 μF, yielding a single peak with little change in bandwidth or insertion loss. This circuit will tune over a reasonable range with a dual section variable capacitor.

The filters of **figs. 13** and **14** are still basically lowpass designs, even though they emphasize the bandpass characteristics. The shunt end capacitors, when viewed in light of the bandpass filtering, serve merely to transform the end load resistance to a different value. Similarly, the shunt coupling capacitor between PLPF sections (**fig. 14**) controls the energy in one resonator shared by the other. Similar results would occur if these elements were shunt inductors with reactances equal to the capacitors previously used. These changes were made, resulting in the circuits of **fig. 15.** These are true bandpass filters, showing infinite attenuation at zero frequency.

The circuits of **fig. 15** have two very interesting properties. First, the response shapes are very symmetrical. This symmetry is maintained even if the two-resonator filter is redone for a wider bandwidth. The second feature is the wide tuning range available. Note

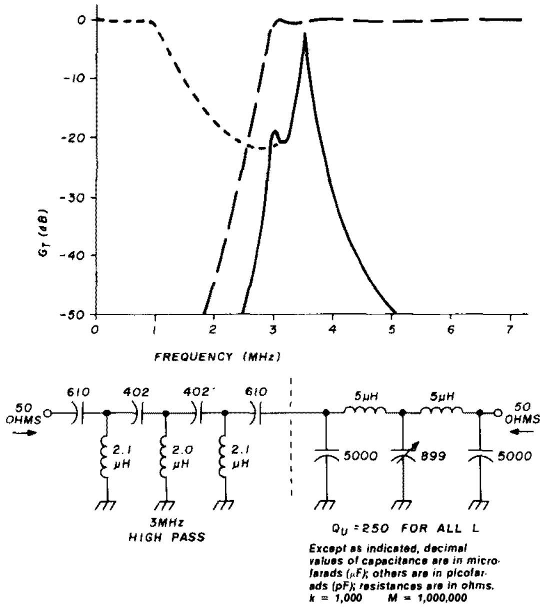


fig. 13. Use of an extreme version of the 5th order PLPF as a bandpass filter. Attenuation at 14 MHz is nearly 100 dB. Dotted and dashed curves show the response of the individual PLPF and the 7th order highpass filter, respectively.

that the only capacitors in the network are those used for tuning. All other elements are inductors. When the same concepts are applied to a more conventional double tuned circuit, the tuning range is expanded. The computer analysis showed that decreasing the two capacitors of **fig. 15B** from 766 to 95 pF produced a peak at 10 MHz, well above the original 3.5 MHz peak.

The double tuned circuit of **fig. 15B** was scaled to a slightly higher frequency and a model was built. Four 3-microhenry inductors were used with a dual section 400 pF variable capacitor. The main inductors were wound on T50-6 toroids, while smaller cores were used for the shunt elements. The circuit was tested with a spectrum analyzer and tracking generator (Tektronix 7L14 and TR-502) to confirm the calculations. This filter was especially easy to align and tuned 6 to 18 MHz. The circuit operates with an approximately constant loaded filter Q — that is, the bandwidth increases as the filter is tuned to higher frequency. Insertion loss varies little with tuning. Agreement with the computer modeling was excellent.

The work presented is restricted to doubly terminated designs. All circuits described have been built, confirming the computer results.

The first step in any filter design should be a careful evaluation of the requirements of the filter. The PLPF is well suited to some applications where a true low pass response is not really needed. The Chebyshev is not the only viable design for a low pass filter.

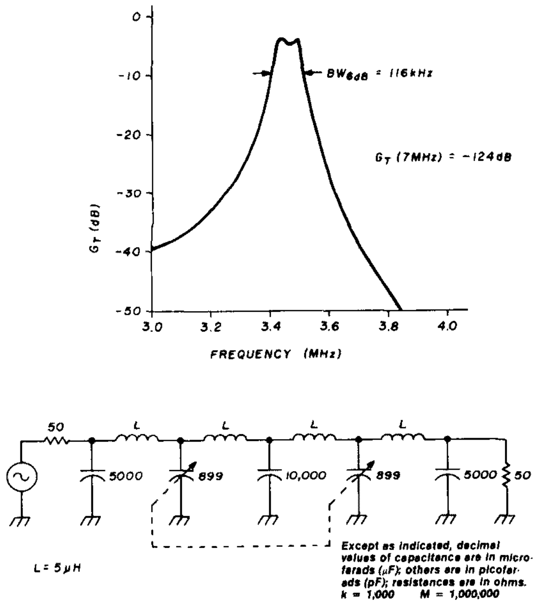


fig. 14. Result of cascading two PLPFs to form a double tuned circuit. See text for discussion on eliminating the double peak, a result of overcoupling.

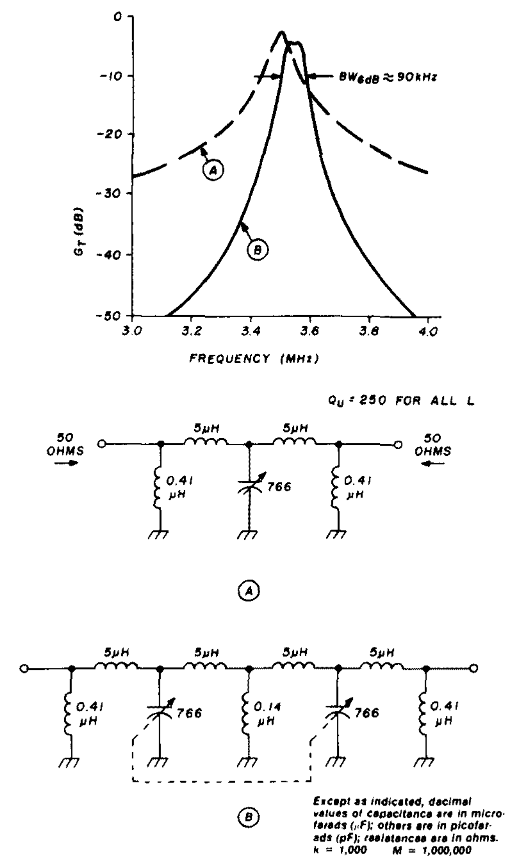


fig. 15. Wide tuning range bandpass filters that evolved from the PLPF.

Computer methods are very powerful, but are no substitute for formal analysis and synthesis. The best computer applications in electronics seem to be those that extend our intuition about circuit design.

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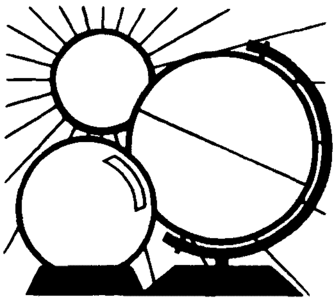


DX FORECASTER

Garth Stonehocker, KØRYW

sporadic E (E_s) DX

Last month the fundamentals of E_s and E_s propagation were reviewed. Now, before the E_s season gets further along, let's look at how E_s propagation can be used for DX. Because E_s propagation is short skip — i.e., 900 miles (1450 km) in hop length — the take-off angle needs to be low (5 to 10 degrees off the ground) in order to obtain the maximum length per hop. This keeps the number of hops (signal loss) to a minimum. For the higher frequency bands where horizontal beam (Yagi) antennas are used, this means tall towers with heights exceeding 60 feet — or even better, 100 feet or more. On the lower frequenices (below 10 MHz) vertical antennas in clear, treeless settings, situated over moist earth and equipped with sufficient ground systems are needed to obtain 5 to 10 degree take-off angles. (To obtain substantial energy at these angles, even at 7 MHz, ground systems should be measured in terms of wavelengths. — **Editor**).

To obtain the highest probability of "reflecting" from an E_s "cloud," a fairly wide beamwidth should be used. Because the beamwidth of Yagis (50 to 60 degrees) is better than the beamwidth of rhombics (20 to 30 degrees), Yagis are preferred for "hitting the clouds." E_s clouds usually measure about 10 by 100 km in length and about 0.6 km in depth. Their thin, dense configuration results in mirror-like reflections rather than the refractions that are characteristic of the F region. Reflection enhances signal strength by an average of 25 dB over refraction.

Another rule of thumb for E_s DX is to use the lowest frequency that isn't absorbed too much. In other words, during daytime, don't use the 10-meter band when 20 is available. The probability of E_s occurrence is higher when the frequency is lower. You can expect an increase of at least 6 to 7 dB in E_s signal strength by using the lowest band that is open.

last-minute forecast

DX conditions are expected to be best *for the higher frequency bands,* 10 to 30 meters (daytime bands), from the 12th through 21st of the month, providing long and short skip openings. The lower frequency bands, 30 to 40 meters, are expected to be the best the first few days of the month and the last week of the month (the 25th and later), including some daytime openings when the solar flux is 80 units or below. The 80 to 160 meter bands will be poor because of noise, except for some sporadic E short-skip openings toward the end of the month.

The Aquarid meteor shower starts about the 18th, peaks about the 28th, and lasts until about August 7th. The maximum radio-echo rate will be 34 per hour. The full moon is on June 13th, lunar perigee the 7th, and solstice on the 21st at 0502 UT.

band-by-band summary

Six meters will provide occasional openings to South Africa and South America around local noontime by short-skip E_s.

Ten meters will be open to the south and southeast for a short period before local noon; to the south at noon and to the southwest in the afternoon. Openings will last longer when the solar flux is at a maximum and improve (transequatorial one-long-hop) during periods of geomagnetic field-disturbances. Listen to WWV at 18 minutes after the hour for pertinent announcements.

Fifteen and twenty meters, almost always open to some part of the world, will be the main daytime DX bands. *Twenty meters* should stay open on long southern paths into the night, though 15 will drop out in the late afternoon. Operate on 15 first, then move down to 20 meters later. DX is 5,000 to 7,000 miles (8,000 to 11,300 km) on these bands. There may be some one-long-hop transequatorial propagation.

Thirty and forty meters are both daytime and nighttime bands. Intermediate distance operation, 1000-1500 miles (1600-2400 km), in any direction is considered daytime DX. Nighttime DX on these two bands may be expected to occur over greater distances than on 80 meters and, like 80, will follow the darkness path across the sky. Signal strength and distances covered are reduced on days of high solar flux values. In addition, no 30-meter openings will take place during the predawn hours on the morning after these high radio flux values.

Eighty and one-sixty meters will exhibit short skip conditions during daylight hours and lengthen for DX near dark. Eighty meters will open to the east just before your sunset, swing more to the south as midnight approaches, and end up in the Pacific areas during the hour or so before dawn. (160 opens later and ends earlier.)

ham radio

JUNE

WESTERN USA

GMT	PDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	5:00	20	15	15	10	20	10	10	15
0100	6:00	20	15	15	10	20	10	10	15
0200	7:00	20	15	20	10	20	10	10	15
0300	8:00	20	20	20	10	30	10	10	15
0400	9:00	20*	20	20	10	*30*	*10*	*10*	15
0500	10:00	15	20	20	10	30	10	10	15
0800	11:00	15	20	20	15	30	10	15	15
0700	12:00	15	*20*	20	15	30	10	15	15
0800	1:00	15	*20*	*20*	15	30	15	15	20
0900	2:00	20	*20*	*20*	*20*	*30*	15	20	20
1000	3:00	20	*20*	*20*	*20*	*30*	20	20	20
1100	4:00	*20*	20	*15*	20	*30*	*20*	20	*20*
1200	5:00	*20*	20	15	20	*30*	*20*	*20*	*20*
1300	6:00	*20*	20	15	20	*30*	*20*	*20*	*20*
1400	7:00	20	20	15	20	*30*	20	*20*	*20*
1500	8:00	20	20	15	20	*30*	20	20	*20*
1800	9:00	20	20	10	20	20	15	20	20
1700	10:00	20	20	10	20	20	15	20	20
1800	11:00	20	20	10	20	20	15	15	20
1900	12:00	20	20	10	15	20	15	15	20
2000	1:00	20	20	10	15	20	15	15	20
2100	2:00	20	15	10	15	20	15	15	20
2200	3:00	20	15	15	15	20	10	15	20
2300	4:00	20	15	15	10	20	10	10	20

MID USA

GMT	MDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN	CDT
0000	6:00	20	15	15	15	20	10	10	20	7:00
0100	7:00	20	15	20*	20	20	10	10	20	8:00
0200	8:00	20	15	20	20	20	10	10	15	9:00
0300	9:00	20	20	20	20	30	10	10	15	10:00
0400	10:00	*20*	*20*	*20*	*20*	*30*	*10*	*10*	15	11:00
0500	11:00	20	20	20	20	30	10	10	15	12:00
0800	12:00	20	20	20	20	30	10	15	15	1:00
0700	1:00	20	20	20	20	30	15	15	20	2:00
0800	2:00	*20*	*20*	*20*	20	30	15	15	20	3:00
0900	3:00	*20*	*20*	*20*	*20*	*30*	15	15	20	4:00
1000	4:00	*20*	*20*	*20*	*20*	*30*	*20*	20	*20*	5:00
1100	5:00	*20*	*20*	*15*	*15*	*30*	*20*	*20*	*20*	6:00
1200	6:00	*20*	*20*	15	15	*30*	*20*	*20*	*20*	7:00
1300	7:00	*20*	20	15	15	*30*	*20*	*20*	*20*	8:00
1400	8:00	20	20	15	15	*30*	*20*	*20*	*20*	9:00
1500	9:00	20	20	15	10	*30*	20	20	*20*	10:00
1800	10:00	15	20	10	10	20	20	20	20	11:00
1700	11:00	15	20	10	10	20	15	20	20	12:00
1800	12:00	15	20	10	10	20	15	15	20	1:00
1900	1:00	15	20	10	10	20	15	15	20	2:00
2000	2:00	15	20	10	10	20	15	15	20	3:00
2100	3:00	15	15	10	10	20	15	15	20	4:00
2200	4:00	20	15	15	15	20	10	15	20	5:00
2300	5:00	20	15	15	15	20	10	10	20	6:00

EASTERN USA

GMT	EDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE CARIBBEAN S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	6:00	20	20	15	15	20	10	10	20
0100	9:00	20	20	15	20*	20	10	10	20
0200	10:00	20	20	20	20	20	10	10	20
0300	11:00	20	20	20	20	30	10	10	15
0400	12:00	20	20	20	20	30	10	10	15
0500	1:00	20	20	20	20	30	10	15	15
0800	2:00	20	20	20	20	30	10	15	15
0700	3:00	*20*	20	20	20	30	15	15	20
0800	4:00	*20*	*20*	20	*20*	30	15	15	20
0900	5:00	*20*	*20*	*20*	*20*	*30*	15	*20*	*20*
1000	6:00	*20*	*20*	*20*	*20*	*30*	*20*	*20*	*20*
1100	7:00	*20*	*20*	*15*	*15*	*30*	*20*	*20*	*20*
1200	8:00	*20*	20	15	*15*	*30*	*20*	*20*	*20*
1300	9:00	20	20	15	15	*30*	*20*	*20*	*20*
1400	10:00	20	20	15	15	*30*	*20*	*20*	*20*
1500	11:00	20*	20	10	10	*30*	20	20	*20*
1800	12:00	15	20	10	10	20	20	20	20
1700	1:00	15	20	10	10	20	15	20	20
1800	2:00	15	20	10	10	20	15	15	20
1900	3:00	15	20	10	10	20	15	15	20
2000	4:00	15	20	10	10	20	15	15	20
2100	5:00	15	20	10	10	20	15	15	20
2200	6:00	20	20	15	15	20	10	15	20
2300	7:00	20	20	15	15	20	10	10	20

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during 'normal' hours.

*Look at next higher band for possible openings

product

REVIEW

ProSearch PSE-2N computerized antenna control

With all the advances in computer technology and applications, it was only a matter of time before someone designed a computer-controlled antenna rotator. The ProSearch PSE-2N is the first microprocessor controlled unit and features enunciated beam headings (with a computer generated voice). It will scan, is programmable for up to ten different beam headings, and has a computer actuated braking circuit. Like other ProSearch controllers, the PSE-2N is compatible with all commonly available rotators.

When I asked Gary Wurdack, President of ProSearch, for some background on the ProSearch Controller, he explained that the design came from his operating experience at KØVUW, where he is actively involved in many contests, and that the ProSearch controller design evolved from his efforts to simplify and streamline the station. He explained his "wish list" of capabilities to a team of design engineers, and asked them to produce a computer controlled antenna rotator. The result is the ProSearch line of controllers.

Of particular interest to *ham radio* readers who have computer-integrated their stations, the PSE-2N can be hooked up to a master computer so that it can be controlled from a single source. This offers a number of exciting possibilities for complete computerization of ham shacks, and it will be interesting to see what results.

design

The heart of the ProSearch PSE 2 controller is a microprocessor that will process, store, and retrieve information and antenna headings and will control, in a manual mode, the operation of the rotator. Command entries are made through a 16-button keyboard on the front of the unit. Output commands from the microprocessor are fed to three driver amps that will activate the left and right movement of the rotator and the brake function. The microprocessor also controls the "talker" voice signals. When numbers are entered through the keyboard, they are read back for confirmation. The computer also controls the 80-Hz tone for CCW rotation and the 400-Hz tone for CW rotation.

As the rotator turns, a positive feedback potentiometer sends analog voltage signals back to the controller, indicating which direction the antenna is pointed. This signal is filtered and then converted into a digital signal to be compatible with the PSE microprocessor.

One admirable feature of the PSE-2N is its ten storable memory locations. Five of these memories have special buttons, keys 1-5, on the keypad for easy access. These areas are marked "Japan," "Europe," "Africa," "South America," and "New Zealand," respectively. Keys 6-0 can also be used to store beam headings for other areas of the world.

operation

Once you've connected all the rotator wires to the Cinch-Jones plugs and connected power to the unit, you are ready to go.

To turn the unit on, simply push the "SCAN & 7" button. All control instructions are entered from the keyboard and each key stroke is verbally confirmed.

The first order of business after installation is to calibrate the unit. This process is quite simple to accomplish using the step-by-step instructions provided and will take only a few minutes to accomplish. After the unit has been calibrated, it is ready to use in normal operations.

In order to demonstrate how the unit determines a specific heading, let's arbitrarily choose 45 degrees for Europe. Punch in a three-figured azimuth, (045 in this case), and push the "GO" button twice. Before the brake unlocks, the unit will say "0-4-5-GO-GO." The unit will unlock and begin to turn. Because of inaccuracies in rotor potentiometers, the computer will need to sample the analog input several times before stopping the unit in a desired direction, but when it stops, your antenna will be within ± 4 degrees of the desired direction.

ProSearch has built a protective circuit into the PSE-2N that will prevent accidental rotor damage from running the unit at full power into the stops: on north-centered units, inputs are within 15 degrees of the stops above 164 degrees or below 195 degrees; on south-centered units, the stops are between above 344 degrees and below 015 degrees. Should there be a heading you want in one of these areas, you can manually direct the antenna into these areas.

programming

To program the antenna for one of the ten memories, you first enter the beam heading. (Let's choose 045 degrees for Europe again.) Push the "STORE" button, then push the 2/EU (Europe). Memory position 2/EU is now programmed at 045 degrees and can be recalled by simply pushing the "GO" button once and then the 2/EU button. The antenna will automatically turn toward Europe at 045 degrees. Button 1/JA, 3/AFRICA, 4/SOUTH AMERICA and 5/NEW ZEALAND can now all be programmed with the appropriate information for your location. Memories 6-0 also can be programmed with any additional headings that you might desire or require.

Another useful feature is Scan. The PSE-2N has five "scan" functions. Scan-1 will have the rotator scan from 0-90 degrees; Scan-2, 90-180 degrees; Scan-3, 180-270 degrees; Scan-4, 270-360 degrees; and Scan-5, a full 360 degrees.

When in the Scan mode the antenna will swing back and forth between the desired section while you look for the station and the optimal signal. Once you hear the station and you get the maximum signal strength, hit any key. The rotator will stop, locked onto the signal.

product REVIEW

While scanning is a pleasurable activity, ProSearch does not recommend protracted amounts of scanning. They suggest that you don't scan for more than ten minutes without giving your rotator at least a few minutes' rest. This is required so that the brake solenoid will have sufficient time to cool.

The ProSearch PSE-2N isn't just a fully automatic unit. As you might expect, you can also turn the antenna manually with a clockwise or counter-clockwise rotation. This is helpful in fine tuning the direction of the beam to maximize signals.

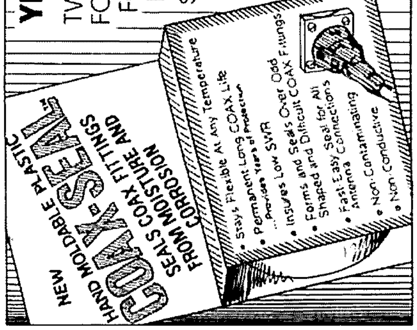
computerization

Here's an area where you can really be at the forefront of current technology. When engineers designed the PSE-2N, they were able to provide an interface port that will facilitate connecting the controller to your home computer. The possibilities that this opens up are truly exciting. One thought is that you could design your station to be run completely by remote control. Why, you could even program a computer to work DX stations while you're at work or at home, sound asleep! You could store all the pertinent information about a particular station and your station could automatically turn on, aim the beam in the right direction, select the appropriate antenna and . . . well it *is* possible and may, in fact, already have been done. It is a fascinating thought; anyone who's actually done it is invited to drop us a note and tell us all about it.

summary

When I write a product review, I generally take the unit home to put it through its paces. In this case, because of the unique nature of the unit, I installed it into our club station WB1AHV so that all the hams here could try it out. It was interesting to hear the first impressions of the staff and station visitors to the PSE-2N; all were pleasantly surprised by the talk-back feature, and commented that it's a valuable asset. In casual operating, all who used the PSE-2N found it to be well engineered and easy to use, even with a minimum of instruction. One of the real benefits I found was its simplified operation. Instead of pushing and holding down both the brake and rotator control, insertion of desired heading plus the execute command left my hands free to do other things. In a contest this could free up valuable time for copying, logging, or checking QSO rate and multiplying counts. The ProSearch PSE-2N was a real "fun" piece of equipment to review.

The ProSearch line of rotator controllers uses high quality components throughout, and all double-sided glass board to ensure long, trouble-free service. The PSE-2N is priced at $469. The PSE-1A "Contester," which sells for $229.95, provides many of the same features, such as memory and digital input, but without "talk



back" and computer interface capability. ProSearch is making a name for itself with these controllers — any one of which would be a valuable addition to any ham shack.

If you'd like more information, contact ProSearch at 1344 Baur Boulevard, St. Louis, Missouri 63132.

Circle #303 on Reader Service Card.

N1ACH

NEW products

GaAs FET receiver preamp

Hamtronics, Inc. has just announced a new low-noise preamp that uses a new dual-gate GaAs FET recently designed especially for service in the VHF/UHF bands. Up until now, to get the low-noise figure of a GaAs FET, a designer had to adapt a transistor originally intended for microwave service; while such transistors work well, they are often costly and tend to oscillate because they have so much gain at the lower VHF and UHF frequencies. As single gate devices, they tend to have the characteristically high feedback capacitance associated with triodes, which makes them hard to tame under a wide variation in load impedances.

The new LNG-( ) series of preamps provides good gain and moderately low noise figure (0.7 to 0.8 dB, depending on band), at a cost of $49, compared to anywhere from $80 to $125 for the earlier type of GaAs FET preamps.

The units operate on standard 12 to 14 VDC, are easy to tune, and may be mounted anywhere, including the tops of towers. LNG preamps are available for all Amateur bands, from 10 meters through 450 MHz.

For more information, including a free catalog on other Hamtronics® products, contact Hamtronics, Inc., 65 Moul Road, Hilton, New York 14468-9535.

Circle #302 on Reader Service Card.

tool and instrument catalog

Contact East is offering a free copy of its 1984 Electronic Tool and Test Instrument catalog, featuring over 5,000 technical products for assembling, testing, and repairing electronic equipment. Products listed include precision hand tools, test instruments, tool kits, soldering supplies, plus a full selection of static control products. All products are fully illustrated with photographs accompanied by detailed descriptions, and prices.

For a copy, write to Contact East, 7 Cypress Drive, P.O. Box 160, Burlington, Massachusetts 01803.

Circle #301 on Reader Service Card.

flea market

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DEADLINE 15th of second preceding month.

SEND MATERIAL TO: Flea Market, Ham Radio, Greenville, N. H. 03048.

QSLs & RUBBER STAMPS — Top Quality! Card Samples and Stamp info — 50¢ — Ebbert Graphics 5R, Box 70, Westerville, Ohio 43081.

WANTED: CW transceiver low band coverage, good working condition and fair price. (518) 883-5007.

WANTED: Old RCA, Western Electric, tubes, speakers, amps. (713) 728-4343, Corb, 11122 Atwell, Houston, Texas 77096.

WANTED: Schematic for Stoner model Pro-10. E.H. Newman, 4873 Nelson St., Fremont, CA (415) 657-1322.

TRAVEL-PAK QSL KIT — Converts post cards, photos to QSLs. Stamp brings circular. Samco, Box 203-c, Wynantskill, New York 12198.

RADIO ITEMS before 1930 wanted. Buying battery operated radios, horn and cone speakers, radio tubes and parts, radio literature — books, catalogs, magazines, radio advertising signs, posters. Gary Schneider, 6848 Commonwealth Blvd., Parma Heights, Ohio 44130.

ELECTRONIC CMOS Keyer kit pcb + parts only $9.95 plus $1.50 shipping. WI res. add 5% tax. Send for free information. BEL-TEK, PO Box 125H, Beloit, WI 53511.

WANTED: ASAP FC-102 antenna tuner with or without FAS-1-4R antenna selector unit. Will consider trade. KA1PSQ (212) 729-0714 collect.

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NOTICE: Buy, sell, trade Amateur Radio, computers, video equipment on our Computer Bulletin Board seen worldwide. Free access to all. Large SASE for details. Back in service May 1984. Narwid Electronics, 61 Bellot Road, Ringwood, NJ 07456.

RUBBER STAMPS: 3 lines $4.50 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

"THE SWAP LIST" has bargains galore. Subscribe now! 6 months for $4.00; 1 year only $6.50. The Swap List, Box 988-H, Evergreen, CO 80439.

ATTENTION C-64 users: Don't buy a logging program until you've read our fact sheet. For free information, write to Crumtronics, PO Box 6187, Ft. Wayne, IN 46896.

REPAIR, ALIGNMENT, calibration. Collins written estimates $25; non Collins $50. K1MAN (207) 495-2215.

FOR SALE Regency HR 312 2m transceiver with mag mount GAM antenna and mike $90.00. Perfect condition. Dentron monitor tuner 2 kw perfect $100.00. W1CPI (401) 789-1817.

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WANTED: Cash paid for used speed radar equipment. Write or call: Brian R. Esterman, PO Box 8141, Northfield, Illinois 60093. (312) 251-8901.

FOUR U.S. plate blocks only one dollar when you request approvals. Please include phone number for reference. Littleflower Stamp Company, RR1 - Box 118(U). Osceola, WI 54020.

STERLING SILVER callsign jewelry pin or tack; $11.95, alligator tie clip; $19.95, all ppd. Info-SASE. Tom's Silver, PO Box 3758, Manchester, NH 03105.

TEKTRONIX scope, single trace, .5 MHz, model 503, $85. Model 516 dual trace, 15 MHz, $200. Plugins, CA $75, 3T77A $75. Tek miniscope model 221 5 MHz single trace, just calibrated, great for field service use, has internal Nicads, carrying case, service manual $700. All postpaid. Dave Roscoe, W1DWZ, 49 Cedar, East Bridgewater, MA 02333. (617) 378-3619.

RECONDITIONED TEST EQUIPMENT $1.00 for catalog. Walter, 2697 Nickel, San Pablo, CA 94806.

FOX-TANGO Newsletters — Since 1972, the prime source of modifications, improvements, and repair of Yaesu gear, free to Club members. Calendar year dues still only $8 U.S., $9 Canada, $12 elsewhere. Includes five year cumulative index by model numbers, or send $1 for index and sample Newsletter. Fox Tango Club, Box 15944, W. Palm Beach, FL 33416.

RTTY-EXCLUSIVELY for the Amateur Teleprinter. One year $7.00. Beginners **RTTY** Handbook $8.00 includes journal index. P.O. Box RY, Cardiff, CA 92007.

IMRA International Mission Radio Assn. helps missioners — equipment loaned; weekday net, 14.280 MHz, 2-3 PM Eastern. Br. Frey, 1 Pryer Manor Rd., Larchmont, NY 10538.

"HAMS FOR CHRIST." Reach other Hams with a gospel tract sure to please. Clyde Stanfield, WA6HEG, 1570 N. Albright, Upland, CA 91786.

TENNATEST — Antenna noise bridge — out-performs others, accurate, costs less, satisfaction guaranteed. Send stamp for details, W8URR, 1025 Wildwood Road, Quincy, MI 49082.

WANTED: Early Hallicrafter "Skyriders" and "Super Skyriders" with silver panels, also "Skyrider Commercial", early transmitters such as HT-1, HT-2, HT-8, and other Hallicrafter gear, parts, accessories, manuals. Chuck Dachis, WD5EOG, The Hallicrafter Collector, 4500 Russell Drive, Austin, Texas 78745.

HAVE A-M CAPABILITY? Join S.P.A.M. (Society for Promotion A-M). Membership is free. Write S.P.A.M. c/o F. Dunlap, 14113 Stoneshire, Houston, TX 77060.

VERY in-ter-est-ing! Next 4 issues $2. Ham Trader "Yellow Sheets". POB356, Wheaton, IL 60189

QUALITY COMPONENTS Transco #11300 UHF relays dual 28VDC coils milspec new $80.00. FXR UHF relay 26VDC coil hermetic milspec new $45.00. Collins filter F455FA15 $65.00. High voltage diode stacks 15kV one ampere milspec new $9.25. Power solenoids 3PST 120A new $45.00; 50A new $35.00 28VDC coils. Carborundum #887 100 ohm 50 watt noninductive resistors new $9.75. Prices prepaid insured. W6FR (714) 871-3607.

NEW 1.2 GHz antennas from NCG. Base or repeater use: (includes mounting hardware) #1260 - 17 step, 1/2 wave collinear, 10.2 dBi gain, VSWR less than 1.5-1, 100 watts, 50 ohm coax. $145.00. #1217 - 17 step, 1/2 wave collinear, 10.8 dBi gain VSWR less than 1.5-1, 30 watts, 50 ohm coax $85.00. Mobile antenna: #1270M - 10 step, 1/2 wave collinear, 8.5 dBi gain, VSWR less than 1.7-1, 20 watts, type "N" connector $54.50. For more information contact: NCG, 1275 North Grove St., Anaheim, CA 92806.

DIGITAL automatic displays for FT-101's, TS-520s, and most others. Six 1/2" digits. Write for information. Grand Systems, PO Box 2171, Blaine, Washington 98230. (604) 530-4551.

WANTED: Schematic for Gonset GR-211 receiver. W.J. Tompkins, 8825 N. Pelham Parkway, Bayside, WI 53217.

RTTY Complete RTTY send/receive station. Printer/keyboard and terminal unit. Nothing else to buy. $200. I ship. H. Parks, 24 Caryl Ave., 6-C, Yonkers, NY 10705. (914) 963-0689.

FOR SALE: Tempo One ac power supply with D-104 mike. Digital two frq. counter extra power supply $400. Write W. Gabali, Rt. 8, Box 624, Lexington, SC 29072.

ATTENTION KENWOOD and Icom owners. Informative separate newsletters. 5th year of publication. Back issues available. Add more selectivity to your 430, 830, and 930. Magicom RF clipper processor for TS-120, 130, 430, 520 and 820. Call for prices. IC-730 FM kit $79.95. TS-830S FM kit $94.95. Send SASE (37¢) for free brochure to International Radio, Inc., 364 Kilpatrick Ave., Port St. Lucie, FL 33452 (305) 335-5545. Master/Visa accepted.

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FOR SALE: B&K model E-200D RF signal generator $280.00; electronic navigation model 310L 250kH-110 MHz RF linear amplifier $500.00; Leader model LSG-231 FM stereo signal generator $300.00; Eico model 150 solid state signal tracer $100.00; Lafayette Pip-Speak stereo speakers $50.00; Leader VLM model LP-70B 20k ohm per volt $40.00; Minolta SRT-200 35mm camera with UV filter, 50mm lens and case $125.00; RF Power Labs model V350 2 meter 350W linear amp. with fan cool $800.00; Monroe simplex autopatch, commercial quality, with various remote functions with Yaesu CPU-2500 2 meter memorizer radio $1250.00. Mitchell Rakoff, 114-41 Queens Blvd., Suite 148, Forest Hills, NY 11375 (212) 591-0002.

Coming Events

ACTIVITIES

"Places to go..."

CALIFORNIA: The Satellite ARC's 1984 Santa Maria Swapfest and Barbecue on Father's Day, June 17 at the Union Oil Company Picnic Grounds south of Santa Maria. General Admission 9 AM. Barbecue served at 1 PM. For information, tickets or swap table reservations: Satellite ARC Swapfest, PO Box 5117, Vandenberg Air Force Base, CA 93437.

GEORGIA: The Atlanta Hamfestival 1984, sponsored by the Atlanta Radio Club, June 16 and 17, at the Atlanta Civic Center. 70,000 square feet of air-conditioned exhibitor space and over 800 outdoor flea market spaces will be available. Flea Market $12.50 per space in advance; $15.00 at the gate for both days. Hamfest registration $5.00 in advance, $6.00 at the door. To be pre-registered for the Flea Market or Hamfest, we must receive your application and check by June 8. Pre-registration applications received after June 8 will be returned. Hours 8 AM to 5 PM on Saturday. 8 AM to 2:30 PM on Sunday. Talk in on 3.97 MHz, 146.22/82 and 146.94 simplex. For pre-registration or other information write Atlanta Radio Club, PO Box 77171, Atlanta, GA 30357.

ILLINOIS: "RADIOFEST '84" a display and sale of antique and classic Amateur equipment as well as vintage radio memorabilia, June 21-23, Holiday Inn, I-90 and Illinois 31, Elgin. This event is sponsored by the Antique Radio Club of America and hosted by the Antique Radio Club of Illinois. Amateur radio participation is welcomed. Talk in on 146.52. For details write Joe Willis, Box 14732, Chicago, IL 60614.

INDIANA: State ARRL Convention and Indianapolis Hamfest, Saturday and Sunday, July 7 and 8, Marion County Fairgrounds, I-74 and 465. Flea Market setup 8 AM July 7. Free camping with hookups available on grounds. Nearby motels. Commercial building open to public 8 AM Sunday. Tech forums all day. Food service. Tickets $4.00 includes free parking and all activities. For further information: Indianapolis Hamfest, Box 11086, Indianapolis, IN 46201.

INDIANA: The Lake County Amateur Radio Club will hold its 12th annual "Dad's Day" Hamfest, June 17, Industrial Arts Building, Lake County Fairgrounds, Crown Point. 8 AM to 2 PM. Tickets $2.50. Plenty of parking and food. Talk in on

147.84/24 or 52. For further information: Bill De Geer, W9TY, Chairman, 3601 Tyler Street, Gary, IN 46408.

MARYLAND: The Frederick Amateur Radio Club's 7th annual Hamfest, June 17, Frederick Fairgrounds. 8 AM to 4 PM. Admission $3.00. YL's and children free. Tailgaters $2.00 additional. Exhibitors tables $10.00 each; $5.00 each additional table. Gates open for exhibitors 8 PM June 16. Overnight security. For additional information: Jim Devilbiss, WA3FUJ, 915 Pine Avenue Frederick, MD 21701. (301) 662-5784.

MICHIGAN: The Straits Area ARC's annual Swap-Shop and Computer demonstration at Emmet County Fairgrounds 4-H Building, Petoskey, July 21. 9 AM to 2 PM. Admission $2.50. Tables $3.00 with setup at 8 AM. RV camping nearby. Talk in on 146.67 and 52. For information: Irene Stein, KA8NKS, 4487 Robinson Rd., Pellston, MI 49769. (616) 539-8986.

MICHIGAN: The Independent Repeater Association of Grand Rapids will hold its annual Hamfestival, Saturday, June 30, 8 AM to 4 PM, Wyoming National Guard Armory, 44th Street, west of US-131. Free table space to all sellers. Admission $3.50. Satellite operation, packet radio, W5LFL space shuttle movie, Amtor forum, CW RX contest, Antenna forum and shack picture contest. Large swap area. Talk in on 147.165/147.765. For information: Linda Hurley, WD8OHW (616) 457-1253 or write I.R.A., 562 - 92nd Street SE, Byron Center, MI 49315.

NEVADA: The YL International SSB System's annual convention, June 21-24, Sahara Hotel, Las Vegas. Deluxe accommodations and RV parking at reasonable rates. Activities include a tour of Hoover Dam, Lake Mead cruise, gala stage show, cocktail party, banquet and breakfast buffet. DX forum and business meetings. YLRL ladies are invited to meet Thursday evening at 8 PM. A convention station will operate on 14,332 kHz. For complete details and registration packet send business SASE with 37¢ in stamps to: Jan Weaver, N7YL, 2195 East Camero Avenue, Las Vegas, NV 89123.

NEW JERSEY: The Raritan Valley Radio Club's 13th annual Hamfest, Saturday, June 16, Columbia Park, Dunellen. Gates open 8:30 AM. Lookers $2.00. Sellers spots $5.00 each, own tables. Refreshments available. Talk in on Club repeater, W2QW/R 146.025/.625 and 146.52 simplex. Advance tickets may be purchased from any club member. For further information call Jack, W2IWK (201) 756-2546 or Ted, WB2TKU (201) 725-3481 between 10 AM and 10 PM.

NEW JERSEY: The Jersey Shore Chaverim is sponsoring the third annual Ham & Computerfest, June 10, 9 AM to 4 PM, Jewish Community Center, 100 Grant Avenue, Deal. 7300 sq. ft. of indoor space. Admission $3 per person (children under 12 and XYL's free). Refreshments available. Indoor table $8 and tailgating $3.50 per space. For reserved space SASE with advance payment to Jersey Shore Hamfest, PO Box 192, West Long Branch, NJ 07764 by June 1. Talk in on 147.045 + .6, 145.110 + .6 and 146.52 simplex. Deal, NJ is less than 50 miles from NYC and 70 miles from Philadelphia. For information call Arnold, W2GDS (201) 222-3009.

NEW YORK: The Putnam Emergency Amateur Repeater League (PEARL) will have its 3rd annual Hamfest, Saturday, July 7, 9 AM to 4 PM, St. John's School, Monsignor O'Brien Blvd., Mahopac. General admission $1.00. Indoor tables $5.00 each. Outdoor tailgating $4.00. For advance registration and information: Frank Konecnik, WB2PTP, RD1, 244 C, Carmel, New York 10512. Talk in on 144.535/145.135 and 146.52.

NORTH DAKOTA/MANITOBA: The 21st annual international Hamfest, July 14 and 15, at the International Peace Garden between Dunseith, ND and Boissevain, Manitoba. Tranmitter hunts, mobile judging, CW contest. Excellent camping. For more information: WDØEMY or WDØDAJ, Box H, Dickinson, ND 58601.

OHIO: The Tusco Amateur Radio Club, W8ZX, and the Canton Amateur Radio Club, W8AL, will hold the 10th annual Hall of Fame Hamfest, July 15, Nimishillen Grange, 6461 Easton Street, Louisville. Admission $2.50 advance and $3.00 at gate. Flea Market additional $2.00 per vehicle. Reserved tables available. Mobile checkin on 146.52/52 and 147.72/12. Call W8ZX or W8AL. For reservations/information: WA8SHP, Butch Lebold, 10877 Hazelview Avenue, Alliance, Ohio 44601. (216) 821-8794.

OHIO: The 20th annual Wood County Ham-A-Rama, Sunday, July 8, Wood County Fairgrounds, Bowling Green. Gates open 8 AM. Free admission and parking. Trunk sales. Food available. Advance table rentals $5.00 (dealers only). Saturday setup until 8 PM. K8TIH talk in on .52. For information or dealer rentals SASE to: Wood Co. ARC, c/o Craig Henderson, Box 366, Luckey, OH 43443.

OREGON: The 9th annual Lane County Ham Fair, July 21 and 22, Oregon National Guard Armory, 2515 Centennial, Eugene. Doors open 8 AM both days. Computer demos, tech seminars, swap tables, kiddie korner, snack bar, free parking for RV's — no hookups. Saturday potluck supper. Tickets and swap tables $5.00 each. FCC exams. Talk in 146.28/88, 147.86/26 and 52. For tickets/tables: Tom Temby, Treas., 3227 Crocker Rd., Eugene, OR 97404. Make checks payable to Lane County Ham Fair.

PENNSYLVANIA: The 13th annual Hamfest sponsored by the Milton ARC, Sunday, June 10, rain or shine, Winfield Fire Co. grounds, Rt. 15, south of Lewisburg. 8 AM to 5 PM. Covered spaces available. Registration $3.00. Spouse and kids free. Flea market, auction and contests. Talk in on 146.37/.97 and 146.025/.625. For further details: Jerry Williamson, WA3SXQ, 10 Old Farm Lane, Milton, PA 17847. (717) 742-3027.

PENNSYLVANIA: The annual "Firecracker" Hamfest, Wednesday, July 4, sponsored by the Harrisburg Radio Amateur Club, Bressler F.C. picnic grounds, exit 1 off I-283. Admission $3.00. XYL and children free. Free tailgating. Nearby motels and restaurants. Plenty of parking. Shaded tables. For details/table reservations: Dave, KC3MG, 131 Livingston Street, Swatara, PA 17113 (717) 039-4957.

WEST VIRGINIA: Wheeling Hamfest, Sunday, July 22, Wheeling Park. Flea market, auction, dealers welcome. Under roof tables available. Admission $3.00. For information/reservations: TSRAC, Box 240, RD 1, Adena, OH 43901. (614) 546-3930

WISCONSIN: The South Milwaukee ARC's annual Swapfest, Saturday, July 7, American Legion Post #434, 9327 South Shepard Avenue, Oak Creek. 7 AM to 5 PM. Picnic area, refreshments available on grounds, free overnight camping. Admission $3.00 per person includes "Happy Hour" with free beverages. Talk in on 146.94 MHz FM. For details the club at PO Box 102, South Milwaukee, WI 53172.

ONTARIO: The tenth annual Ontario Hamfest, July 14, 7 AM to 4 PM, Milton, Ontario, fairgrounds. Weekend camping, free parking, free flea market tables. Tickets $2.50 advance, $4.00 at gate. Commercial displays, refreshments. Talk in on Club repeater 21/81. For details and pre-registration: Burlington ARC, PO Box 836, Burlington, Ontario L7R 3Y7.

BRITISH COLUMBIA: The Maple Ridge ARC is hosting Hamfest '84, June 30 and July 1, Maple Ridge Fairgrounds 30 miles east of Vancouver. Registration: Hams $5.00, non-Hams over 12 $2.00. Swap and shop, commercial displays, bunny hunts, ladies' and children's programs and more. Camper space available with elec. Talk-in on 146.20/80 and 146.34/94. For information an registration (20% off gate entrance) contact: Maple Ridge ARC, Box 292, Maple Ridge, BC V2X 7G2

OPERATING EVENTS

"Things to do..."

JUNE 16: The Missouri Valley ARC's fifth annual Pony Express Day. 0900 to 1700 CST. and June 17, 1000 to 1300 CST to commemorate the original running of the Pony Express from St. Joseph, MO to Sacramento, CA. Listen for club station WØNH 10 kc's from bottom of General phone bands on 15, 20, 40 and 75 meters. 10 meters 28.575. CW 28.150 on 10; 21.150 on 15; and 7.125 on 40. Send one 1st class stamp and QSL to Missouri Valley ARC, 401 N. 12th Street, St. Joseph, MO 64501.

JULY 7-8 SSB and JULY 28-29 CW. Venezuelan Independence Worldwide Contest. 0000 GMT Saturday — 2400 GMT Sunday. All bands exchange RS(T) plus a three figure QSO number starting with 001. Logs must show date, hour (GMT only), station worked, reports exchanged and respective numerical order, multipliers and points. Each participant will accompany log with US $2.00 or IRC equivalent postmarked no later than August 15, 1984 for SSB participant and September 15, 1984 for CW. Send logs to RCV, PO Box 2285, Caracas 1010-A, Venezuela.

NEW AMATEUR OPERATING CERTIFICATE now being offered by the Bartlesville (Oklahoma) ARC to focus attention on the "Green Country" region of northeast Oklahoma. This award is available to anyone making two-way Amateur radio contact with three hams in the Nowata, Osage and/or Washington Counties of Oklahoma. All bands/modes permitted. Applicants for the award should submit calls and pertinent details of three qualifying OSO's plus $1.00 s&h to W5NS Awards Manager, 1800 Moonlight Drive, Bartlesville, OK 74006.

THE LINCOLN (NEBRASKA) COMMUNICATIONS SOCIETY has constructed a beacon transmitter to provide a signal for propagation studies and frequency reference. The beacon operates CW on 144.055 MHz and is located in the northeast corner of grid square EN-10. I.D. callsign is WBØQIY/B. Reception reports should be sent to Lincoln Communications Society, Att: KØNG, 1801 So. 48th Street, Lincoln, NE 68506.

MAY-SEPTEMBER: N.O.A.R.S. and the U.S.S. COD will be on the air again during the summer of 1984. NOARS members will operate from the COD starting Memorial Day weekend daily through Labor Day weekend. Look for operations in the lower portion of the General bands 10 through 80 meters. Special Novice operations on June 16, July 15 and August 18. Extra operations during Cleveland Hamfest, September 23. For a special 8 × 11 certificate picturing the U.S.S. COD send QSL confirming two-way contact and $1.00 s&h to WD8RZG.

JUNE 11-17: The Henry County ARC will operate club station K8TII to commemorate the Napoleon, Ohio, Sesquicentennial. Frequencies: 3740, 3965, 7065, 14265, 21150 and 21365. Contact with club station or any club members' stations qualifies for certificate. SASE to Roger C. Jaqua, W8SMW, 17136 Mercer Rd., Bowling Green, Ohio 43402.

JUNE 8-10: The Macomb Emergency Communications Association will have its second special event. Operation commences at 2200Z Friday to 2200Z Sunday. Lower end of General class portion of each Amateur band. SSB and CW/RTTY of HF. FM phone on 146.07/67. QSL to MECA, Box 488, Utica, MI 48087 with 9 × 12 SASE. DX stations need send only QSL.

JUNE 29-JULY 1: The Muscle Shoals ARC will operate W4JNB from 1600-2100Z from Spring Park, Tuscumbia, Alabama to celebrate Helen Keller Festival Days. Phone frequencies: 7270-7290 and 14,280-14,295. For certificate send 4 × 10 SASE to Box 2745, Muscle Shoals, AL 35662.

JUNE 30-JULY 1: The Hannibal ARC will issue a fourth annual special certificate from the National Tom Sawyer Days celebration in Mark Twain's boyhood home town, Hannibal, MO. 1500-21000 UTC both days. Frequencies: Phone — 7.245, 14.290, 21.400, 28.770 and CW — 7.125 and 21.125. To receive the certificate send a large SASE and your QSL card confirming contact to Hannibal ARC, WØKEM, 2108 Orchard Avenue, Hannibal, MO 63401.

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focus

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contents

REFLECTIONS

THE NUMBER 1 QUESTION

I've recently discovered that if I connect two framastans in parallel, water-cool them, operate them in Class E, and drive them with my 2N5109 solid-state QRP transmitter, I can achieve the legal output power level and cover 160 meters through 40,000 GHz without retuning. Would you be interested in seeing a manuscript on this subject?

Every time I read a question like this from a prospective author my interest is definitely aroused. Editors constantly receive telephone and mail inquiries about possible articles covering a wide range of subjects. We at *ham radio* are no exception, because every day we receive proposals for manuscripts ranging in size from quarter-page ham notes through six-part series of epic proportions. In order to encourage you to contribute your wisdom and experience to the pages of *ham radio*, I'd like to tell you what I'd like to see coming through the transom.

If you'll look at a magazine — any magazine — you'll notice that there are a number of realities associated with it. One is the number of pages allocated to editorial content. Ah, but what's editorial content? you ask. Basically, "editorial" content is anything that isn't advertising — technical articles, ham notes, columns, letters, *Presstop*, new product announcements, and even the front cover, as well as (ahem) this editorial. If you were to view each page as if it were a parcel of extremely valuable land, you'd understand the dilemma (and the joy) an editor faces when reviewing manuscripts. So many good ideas . . . and so little space!

We have a mandate from our readers, consistent with our charter and clearly reaffirmed by the Reader Survey of September, 1983, that directs us to provide high quality *technical* material. For my part, I'm more than willing to meet this demand. For those of you who are thinking of proposing an idea for an article, please keep the following guidelines in mind:

Length. Don't worry about it. With very few exceptions (and this editorial may be one of them), the subject will dictate the length of the article. It's difficult to write more than a paragraph or two about how to insert 6146s into sockets (with a hammer, of course), so if you're going to write on a small, limited topic, keep your article short. But if you're going to deal with a larger topic in a more comprehensive way, then give your subject the room it needs; consider, for example, K2BT's series on vertical phased arrays, which ran in six installments spread over a year's time.

Style. Yours is *fine*. Just keep it simple; the point is, after all, to convey information. Don't worry about grammar; we'll worry about that. It may help to remember this rule of thumb: *write to inform, not impress.*

Accuracy. Here's where we all get into trouble now and then. "In conclusion, my superduper loudenboomer metal noodle, described herein, works on the principle of thermoquality of non-homogeneous turf." It very well might. But please prove it. If you've discovered some fabulous new technique that promises to revolutionize the world's understanding of antennas, propagation, or anything else, we want to hear about it . . . but *please* present your idea in a logical sequence of facts and build your case. If you decide to compare your method to a standard, make sure your setup is carefully designed and that it generates reproducible results.

Subject. This I leave to you. However, you might want to ask yourself one question before taking pen in hand. How many readers will be interested? Three, three thousand, or thirty thousand? This is especially true in regard to conversions. Not too many readers are interested in converting a BC610 to 11 meters. However, a simple circuit that adds the three WARC bands to an older transceiver will generate substantial interest. How about a short article on receivers, or on some particularly exciting aspect of future technology? I'll be glad to consider your idea. Just pick up the phone or send a one-page outline of your proposed manuscript. (I can't, by the way, consider simultaneous submissions.)

A final word. Please double space your manuscript. Don't hide or crowd your words . . . let them sing out and tell your story. For this — and more — I thank you.

Rich Rosen, K2RR
Editor-in-Chief

presstop de W9JUV

OSCAR 11 IS BACK ON THE AIR AFTER 10 WEEKS OF FRUSTRATING SILENCE. UOSAT OSCAR 11, the University of Surrey's scientific satellite, finally responded to signals from its parent command station May 14 and is now being tested to uncover its post-launch problem.

Launched March 1 From Vandenberg AFB in California, the English Amateur satellite had gone off the air after only three orbits and then staunchly resisted all attempts to bring it back to life. The key to its resuscitation was provided by a crew in Greenland, who used a 100-foot dish to hear signals leaking from its 1.2 GHz command receiver that both confirmed its actual orbital position and provided clues to unlocking the command hangup!

OSCAR 11's Three Beacons Can Now Be Heard Again on 145.825, 435.025, and 2304.5 MHz.

Another Sophisticated Amateur Satellite Could Be In The Offing, with a projected early 1986 launch date. This possible launch opportunity would permit a large payload, at least as heavy as OSCAR 10, opening the door for still another multi-function Amateur bird.

A World-Spanning Set Of Geostationary Amateur Satellites is also currently under consideration by AMSAT. Operating in the 1.3 or 2.3 GHz bands and stationed at suitable intervals around the equator, these satellites could provide a round-the-clock worldwide communications capability far beyond that of present orbiting Amateur satellites. Such a project would be expensive, however, so before proceeding too far with it AMSAT would like to hear expressions of interest and support. Write AMSAT, Box 27, Washington, D.C. 20044.

VOLUNTEER EXAMINER COORDINATORS ARE NOW IN PLACE IN 10 of the 13 areas designated by the FCC. The sixth district now has a VEC, the Greater Los Angeles Amateur Radio Group (GLAARG), a grouping of area clubs; the San Diego Amateur Radio Council (SANDARC), should also have FCC acceptance before this sees print. Two additional VECs are on board in the fourth district, VE-administered exams have already been given in Alaska and the second, eighth and ninth call areas as this went to press, and GLAARG was hoping to start giving exams in the LA area in conjunction with Field Day. This leaves only the first and zero call areas, and the Pacific, without at least one acceptable VEC, though Wayne Green, W2NSD, has told Westlink that he intends to apply for first district VEC through the radio club at his recently established Wayne Green University.

The Proposal To Permit Exam Expense Reimbursement will probably be considered by the Commissioners by mid-July, and if there's no hitch, exam fees could be in place before Labor Day. If so, expect the ARRL to submit its long-awaited VEC proposal very quickly. Just how soon a League-sanctioned program could get under way is questionable, however. The experience of those VECs already up and running is that both the necessary paperwork and the exam process itself are sufficiently complicated to require some formal VE training. Dayton, for example, gives its VEs two three-hour training sessions. The League, however, expects to prepare its VE corps by simply distributing a "training manual!" Another factor to be resolved is whether the League can build an adequate "Chinese Wall" between its VEC role and its Amateur training and training publications activities, a problem that will also be a potential obstacle to any proposal from a Wayne Green enterprise.

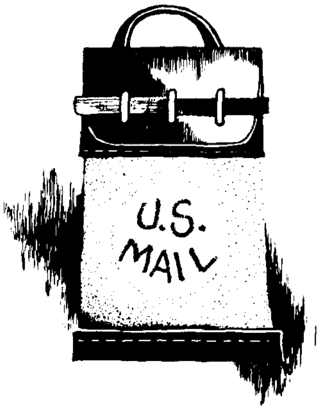
THE 1984 SUMMER OLYMPIC TORCH RUN KICKED OFF MAY 8 in New York City and passed through Chicago two weeks later. The torch bearers each run a four-mile segment, with their efforts coordinated by teams of 16 Amateur volunteers from AT&T's Telephone Pioneers of America. Two meters is used for the bulk of their communications, using both simplex and area repeaters when available. By the time the run ends in Los Angeles in late July, the runners (and their Amateur Radio support teams) will have covered over 9000 miles.

ADDITIONAL EXPANSION OF PHONE BANDS COULD BE DECIDED before the end of July. Unlike the earlier FCC action on phone band expansion, which concerned only 20 meters, this time all other bands in the NPRM—75, 40, 15, and 10 meters—will be considered. However, just which phone bands will be changed and by how much is still an open question. Canada's Department of Communications has proposed doing away with mandated subbands in its Amateur service!

Amateur Access To The New 902 MHz Amateur Allocation may also be relatively imminent. IRAC (the Interdepartmental Radio Advisory Committee) seems to have no objection to giving Amateurs the band on a secondary basis, though RCA is protesting that its video disc players are susceptible to 902 MHz RFI. In the HF spectrum, 24 MHz may also see Amateur occupancy soon as well, but problems with existing users make access to 18 MHz a more distant hope.

Getting New Bands From WARC Has A Negative Aspect, as well. Along with the gains come losses, the first of which will be a severe cut in the 1215-1300 MHz band. Amateurs along the Canadian border will also lose access to 420-430 MHz, and eventually (when AM broadcast expands upward) radiolocation will again lay claim to 1900-2000 MHz.

FURTHER EXPANSION OF 20-KHZ CHANNELIZATION TO TWO METER'S TOP HALF will be a major topic at the Texas VHF-FM Society's August 11 meeting in Austin. Heading up the forum on this controversial change is Joe Jarrett, K5FOG, who seeks inputs from concerned Amateurs across the country. Papers (from those unable to attend in person) and outlines of oral presentations on the subject should go to Joe at 8501 Spring Valley Drive, Austin, Texas 78736.



comments

on the VHF/UHF challenge

Dear HR:

I am especially pleased to see Joe Reisert's VHF/UHF column and have written a personal note of thanks to him. I've been an avid V/UHF'er for over 20 years and try to communicate the excitement and challenge of operating on the VHF/UHF bands to other hams whenever I can. That's why I'm glad to see you giving someone of Joe's stature and expertise a chance to reach a large "lay" audience on a monthly basis!

Jack C. Parker, KCØW
Bismarck, North Dakota

Dear HR:

The January, 1984, issue came today, and I'm very pleased to be able to look forward to Joe Reisert's "VHF/UHF World" every month.

I'm one of a handful of Amateurs in Montgomery who operate 2-meter SSB and plan to be on 432 soon. I am NCS of the Central Alabama 2-meter SSB net each Wednesday night at 0200 (Thursday) UTC on 144.250 MHz.

I also really liked the January "Ham Notebook." There have been many repeater controllers published in various publications, but this is the first time I have seen a repeater-to-phone schematic. This is a circuit I have been looking for — thanks.

Alton L. Erdman, W4CNQ
Montgomery, Alabama

Timex/Sinclair newsletter

Dear HR:

A specialized newsletter is available for Amateurs with Timex-Sinclair computers. *QZX* is a monthly publication that includes programs and news features pertaining to Sinclair computers. At $12 per year, I think it's a great resource for Sinclair computer owners; one recent issue, for example, was dedicated to RTTY applications.

For more information, contact *QZX* at 2025 O'Donnell Drive, Las Cruces, New Mexico 88001.

Tom Hart, AD1B
Westwood, Massachusetts

new MSO for RTTY

Dear HR:

A new MSO (Message Storage Operation) system has been established in the New York metropolitan area by N2CKA. The frequency of operation is 145.680 MHz-FM, using 100 WPM (74 baud) Baudot Code.

All Amateurs in the New York-New Jersey-Connecticut area (for that matter, anyone who can "hit" the MSO from elsewhere) are encouraged to leave items of interest, bulletins, or just "mail."

The MSO uses standard DS 3100 type format. The access code, "MSOCKA.HELP" will bring a list of commands to assist the new user. "EXIT" is used to "close" or deactivate the system.

N2CKA or I will be glad to help users unfamiliar with MSO operation to get off on the right foot.

Rich Tashner, N2EO
Whitestone, New York

can *you* help?

Dear HR:

I am in possession of a 6-channel HF-SSB marine transceiver made by Maritek — Model No. SB6-80, approximate date of manufacture, 1973.

I have been able to trace the company to the following address: 1819 South Central, Kent, Washington, 98031, but letters have been stamped "Not known" and returned. No phone number is listed.

I would like very much to have a copy of the workshop service handbook and/or circuit details, a copy of the operations handbook, and any other information — but specifically that which applies to the settings of the taps of the output stage.

I can arrange to have these photocopied in either the U.S. or Canada, or here in the U.K., and the original returned. I am willing to reimburse any expenses incurred.

Kris Partridge, G8AUU
6 Blagdon Walk,
Teddington, TW11 9LN,
United Kingdom

While the HI MANUALS catalog doesn't list the Maritek SB6-80, it's often a helpful source of hard-to-find manuals and schematics. For a copy, write to HI MANUALS, P.O. Box 802HR, Council Bluffs, Iowa 51502; enclose $1.00.

Editor

operating etiquette

Dear HR:

I'm delighted when editors speak out about the selfish operating practices prevalent on the HF bands (Reflections, *ham radio*, February, 1984). On the accidental occasions when *I* work DX, I want to get to know the other guy and find out interesting things about his country. I would never insult a foreign ham by implying that I am interested in him only for his QSL. If I naively try to converse for more than a couple of minutes, however, the vultures pounce.

I'm also dismayed by hams who hog frequencies, claiming they are participating in phantom "nets" or handling nonexistent "overseas phone-patch traffic." I've heard better operating

practices on the frequencies between 10 and 11 meters! CB isn't to blame; some of the most rules-conscious hams I know are ex-CB'ers who hoped to find something better in Amateur Radio.

These things, plus the obnoxious proliferation of contests, cause me to confine most of my activity to experimental modes of communication where contacts are rare but always stimulating.

There are two kinds of hams — operators and experimenters. Both are necessary, but the operators vastly outnumber the experimenters. Bravo for *ham radio* for supporting us underdogs!

Frank Reid, W9MKV
Bloomington, Indiana

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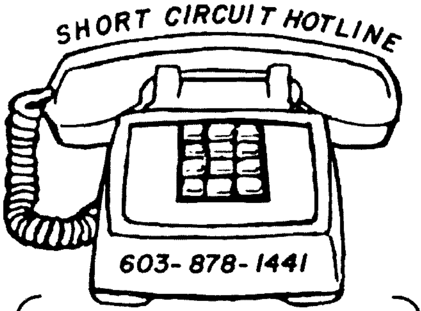
short circuit

filter design

In **fig. 6** of W6NRW's article, "Graphic Filter Design," (April, 1984, page 37), the resistor labeled 528.2 should be labeled 582.2. This nominal 7 percent error in value results in some peaking and a general tilt to the filter's passband.



Building a current *ham radio* project? Call the Short Circuit Hotline any time between 9 AM and Noon, or 1 to 3 PM — Eastern time — *before* you begin construction. We'll let you know of any changes or corrections that should be made to the article describing your project.

(See "Publisher's Log," April, 1984, page 6, for details.)

the VHF/UHF primer: an introduction to propagation

To *really* work DX, try a new mode

One of the most interesting aspects of the world above 10 meters is radio wave propagation. Sure, we've all heard about F2, tropo, meteor scatter, ducting, EME, and perhaps auroral propagation. But are these the prime or only ways of working DX on VHF/UHF? No way!

In preparing my monthly column for *ham radio*, I've noticed that I'll occasionally identify a propagation mode without providing any explanation about how that mode works. Sometimes it's difficult to readily identify a particular mode of propagation on the VHF/UHF frequencies because not all modes are well understood. And different propagation modes are frequently combined, making identification difficult at best.

Over the years, the many articles written about VHF/UHF propagation have only rarely covered more than a few modes at a time. Lesser known modes have been hidden away and almost forgotten. In this article, I'll identify and explain the most common modes of radio propagation, discuss some newly discovered, less common ones, and try to provide some insight into how you can exploit them and perhaps discover some new modes yourself. Wherever possible, references will be listed for those interested in pursuing the individual subjects further.

Before we proceed, remember that communications on any frequency is limited not only by the propagation mode and path length, but by several other factors such as transmitter power, receiver sensitivity, and antenna gain, among others. A full discussion of the latter parameters would fill an entire article, if not a book; hence, I'll limit this discussion to a basic introduction and a review of most of the identified VHF/UHF propagation modes, as well as the operating techniques that make use of these modes. I'll begin with the propagation modes on the lowest VHF frequencies and move on upward through SHF, adding some of my own observations and suggestions about how to use each mode to its best advantage.

line-of-sight

Line-of-sight, surely the most common form of VHF/UHF propagation, is easily understood.[1,2] At one time it was thought to be the *only* reason to use VHF/UHF; even now, the idea of QRM-free local communications is appealing.

Light and radio signals usually travel in relatively straight lines. If the antenna height is known, a simple formula can be used to determine the distance to the horizon:

$$D\ (miles) = \sqrt{1.5H} \qquad (1)$$

where H is in feet,

$$D'\ (km) = \sqrt{12.75H'} \qquad (2)$$

where H' is in meters.

Because of changes in air pressure, temperature, and humidity along the path, the refractive index of the atmosphere may change and bend or refract signals over a greater distance than the normal line-of-sight. Generally speaking, this refractive index averages 1.33, meaning that we can normally communicate over a distance 33 percent further than the normal line-of-sight. To account for this we can modify **eqs. 1** and **2** as follows:

$$D\ (miles) = \sqrt{2H} \qquad (3)$$

or

$$D'\ (km) = \sqrt{17H'} \qquad (4)$$

For example, station "A" has a 50-foot (15.24 meter) tower. It has a horizon of approximately 8.66 miles (13.94 km) and a radio horizon of 10 miles (16.09 km).

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table 1. Claimed 2-meter (and up) terrestrial DX records.

frequency band	record holder	date	mode	DX miles	DX km
144-148 MHz	I4EAT-ZS3B	30 March 1979	Trans. Eq.	4840	7788
220-225 MHz	KP4EOR-LU7DJZ	9 March 1983	Trans. Eq.	3670	5906
420-450 MHz	KD6R-KH6IAA	28 July 1980	Tropo. duct.	2550	4103
1240-1300 MHz	VK5MC-VK7KZ/P	23 Jan. 1980	Tropo. duct.	1422	2290
2300-2450 MHz	VK5QR-VK6WG/P	27 Jan. 1978	Tropo. duct.	1190	1883
3300-3500 MHz	G3LQR-SM6HYG	11 July 1983	Tropo.	576	927
5650-5925 MHz	G3ZEZ-SM6HYG	12 July 1983	Tropo.	610	981
10-10.5 GHz	I0SNY/EA9-I0YLI/IT9	18 July 1983	Tropo. duct.	1034	1663
24-24.5 GHz	DJ2UH/P-DJ4YJ/P	21 Feb. 1982	Tropo.	152	244
48 GHz and up:	none reported				

Note: EME hasn't been included since it would distort the records significantly. EME DX of over 11,000 miles (17,699 km) has been accomplished on 2 meters, 70 cm and 23 cm. 220 MHz is only limited by the lack of frequency allocations worldwide. 13 cm is the highest band where Amateur EME QSOs have been claimed.

Station "B" has a 100-foot (30.48 meter) tower. It has a horizon of approximately 12.25 miles (19.71 km) and a radio horizon of 14.14 miles (22.76 km). If stations "A" and "B" are on opposite ends of a straight line between the same point on the horizon, the distances can be added to find the maximum path length. In this example the line-of-sight distance would be 20.91 miles (33.65 km) and the radio horizon would be 24.14 miles (38.85 km). If two stations are attempting to communicate, any obstructions or trees at the common point on their respective horizons may attenuate signals!

path loss attenuation

Space does not permit me to discuss path loss attenuation at length; see reference 2 for a detailed summary of this phenomenon. In brief, it may be useful to know that typical line-of-sight attenuations for a 10-mile (16 km) unobstructed radio path are 90.5 dB at 50 MHz, 109.3 dB at 432 MHz, 123.8 dB at 2304 MHz and 136.6 dB at 10 GHz. Due to the inverse square law, attenuation increases or decreases by 6 dB every time you either double or half the path length or frequency, respectively. Hence, if you try to double the path length (providing you have line-of-sight), the path loss will be 6 dB greater.

F2 propagation

The supreme propagation mode for DX and the workhorse of the HF bands is F2 propagation. Once it was thought that ultraviolet radiation from the sun would never be sufficient to ionize the ionosphere sufficiently to support communications on 50 MHz via the F2 layer (approximately 150 to 250 miles or 250 to 400 km above the earth), but the experts were proven wrong when W1HDQ contacted G6DH crossband (6 to 10 meters) via this mode on November 26, 1946. Only hours later a transcontinental 6-meter two-way QSO using the same mode between W4GJO in Florida and W6QG in California occurred. Six-meter F2 was not too common during solar cycle 18, and the sun "cooled down" by November, 1947.

The International Geophysical Year (1957-1958) helped numerous stations obtain 6-meter permits where operation was normally forbidden. In October 1957 W4UMF worked SM5CHH, signaling what may have been the start of the greatest F2 openings ever recorded on 6 meters. Later that year K6GDI submitted the QSLs to qualify for the first WAC above 10 meters, showing the worldwide participation. In 1957, solar cycle 19 hit a tremendous peak, the highest recorded in history, when on December 24 to 25, the sunspot count reached 355 and worldwide propagation, even via the long path, became commonplace on 6 meters. But by early 1960, this cycle was on the decline, and unfortunately, solar cycle 20 displayed a more normal solar activity peak level with some 6-meter F2 contacts reported between November, 1967, and February, 1970. Most were single-hop contacts. It began to look as if we'd never see F2 on 6 meters again in our lifetime.

However, in the fall of 1978, during cycle 21, solar activity increased dramatically and F2 propagation began anew on 6 meters. The mean sunspot count in December, 1979, was 164.5, the highest for cycle 21. Unfortunately, many countries no longer permitted Amateur operation on 6 meters because of TV and other frequency allocations. As a result, many VHFers outside the United States had to be satisfied with SWLing or 6 to 10 meter crossband QSOs. 1980 saw a big F2 dip in the United States, but surprisingly, propagation returned for a strong second session in November, 1981, slowly diminishing in late 1982. (As I wrote this column, in March of 1984, F2 returned, with some USA areas working VK's and ZL's. No guess on how long it will last, but enjoy!)

F2 propagation requires a very high sunspot count, typically *greater than* 125, which is roughly equivalent to a 2800 MHz solar flux of 175 (as broadcast at 18 minutes after each hour on radio station WWV). Generally speaking, in the continental United States, openings start first at the southern latitudes, slowly working up to the northern latitudes over a period of days to weeks as the ionosphere gets "pumped up." Hence the solar activity has to be sustained for at least several days in a row. In the continental USA, openings usually occur between late October and March, favoring southern latitude stations.

At the first appearance of F2, single hop distances are right at the MUF and very long (typically 2500 miles or 4000 km), getting shorter as the ionization builds for several days. As the *ionization* increases, the MUF slowly rises, permitting operation above 52 MHz. Six-meter communications with low power (10 watts or less) become possible even with dipole antennas. If conditions remain favorable, multiple hops become possible.

The best propagation between two points usually occurs when the sun is about halfway between the stations and almost always requires that the entire path be in sunlight. At times the path can be rather selective, especially on multiple hops; this means that stations as few as 10 miles (16 km) apart may not be able to hear the same distant station. F2 can also link up with sporadic E (more on this later) and extend the path. When solar disturbances occur, the north-south paths show little or no change and often become enhanced soon after the storm subsides. However, the east-west paths drop off quickly and are usually good only during undisturbed solar periods.

Although several crossband 50-70 MHz QSOs were reported during cycle 21 between VE1ASJ and the United Kingdom, the 70 MHz signals were almost certainly not propagated by F2 but rather by sporadic E, since sporadic E propagation was in evidence on 6 meters during these openings.

Much of the success during cycle 21 can be attributed to improved gear, propagation beacons, monitoring of commercial paging and television stations (such as the BBC on 48.25 MHz), and the various liaison nets, especially those on 28.885 MHz (this frequency is still used whenever 10 meters is open). Most of the 6 to 10 meter crossband operation takes place near this frequency. Solar bulletins on radio station WWV at 18 minutes after each hour are also helpful because they *report solar flux, an* indication of sunspot count, and disturbance warnings via the "A" and "K" indices. Much more information on F2 can be found in "The World Above 50 MHz" column in *QST* during the appropriate years. It's doubtful that any widespread F2 propagation will occur on 6 meters again until about the year 2000, but it sure was great while it lasted!

backscatter

This mode of propagation, a form of radar, often preceeds sporadic E and F2 openings. It is most prevalent on 10, 15, and 20 meters and sometimes occurs even on 40 meters. Its presence indicates that the MUF is very high. In backscatter operation, what happens is that an area of the ionosphere is so highly charged that signals traversing the E or F2 layer path strike the earth at some distant point and are scattered. If two stations running reasonable power and antenna gain aim at this common scattering point, communication is made possible by reflection. Because signals arrive from several *different* directions at the same time (multipath), they are usually weak and distorted. A good time to look for 6-meter backscatter is when propagation is good on 10 meters; even then, it may be difficult to spot unless there is lots of activity. This mode is primarily limited to 6 through 40 meters and can be used for DX even up to 3000 miles (4825 km).

midlatitude intense sporadic E

This is truly the workhorse mode for DX on 6 meters. Propagation is typically from ionized clouds located in the E layer and usually approximately 60 miles (100 km) above the earth in the midlatitudes (25 to 55 degrees north latitude). There is also equatorial and polar sporadic E, but they are slightly different in nature and of little value for continental USA stations since they are too far south or north to be of any use.

In the continental USA, sporadic E propagation usually peaks near the summer solstice (June 21). Typical limits are from May to early August but can occur any time of year. Typical DX limits are about 300 to 1300 miles (485 to 2100 km) per hop but two and even three hops are not uncommon on 6 *meters.* A lesser peak occurs approximately within plus or minus one month of the winter solstice (December 21) but this peak is considerably shorter, weaker and usually limited to single hop. It is worthwhile noting that the peaks also occur in the southern hemisphere in reverse order. Although sporadic E propagation can start at any time of the day, it is most prevalent in the late morning, the late afternoon and into the early evening. There is some speculation that sporadic E occurrences are more common and stronger when sunspot activity is low. Hence propagation via this mode may improve in the next few years.

The exact mechanism that triggers sporadic E is still not known. Some people attribute it to weather-related phenomenon, especially wind shear and lightning storms. The late Mel Wilson, W1DEI/W2BOC, devoted much of his Amateur career to the study of sporadic E and wrote several very interesting articles on the subject.[3,4] He pointed out that the clouds that produce the propagation originate at a location he

called the "birthplace" and usually travel in relatively straight lines at approximately 180 miles (290 km) per hour from southeast to northwest (**fig. 1**). For best propagation, he said, the cloud should be on or nearly on a line drawn between the two stations at the halfway point. Clouds ±5 degrees off the path can be used and up to ± 10 degrees at the extreme. A single cloud will be usable for up to 5 minutes. The size of the cloud has been estimated to be in the order of tens to hundreds of feet (5 to 100 meters) vertically. Multiple clouds are often being formed in the same birthplace so they will pass by as they are generated. The best sporadic E propagation seems to occur when no major high pressure areas are present. (Mel's work will be carried on by his son Steve, W2CAP/1.)

Midlatitude sporadic E is also usable on 2 meters with the first reported contacts made between northern Texas and southern California on July 10, 1951, for a distance of approximately 1300 miles (2090 km). At first, other 2-meter openings were rare, probably due to lack of activity, and weren't recorded again until July, 1956 and June, 1959. Nowadays 2-meter contacts are usually reported yearly and tend to occur between late June and early August, but can occur on the winter peak as described above. Strong lightning activity, especially the type that occurs above 60,000 feet (18,288 meters), is often present near the midpath of big 2-meter openings.[5] Tornadoes and hail storms are often present at the mid-point of the path during 2-meter openings. Typical 2-meter QSOs using sporadic E are shown in **fig. 2** from data given to me by the late Mel Wilson. Although we have had some very long DX (over 1600 miles or 2575 km) 2-meter openings in the United States, I have never heard of a documented double-hop QSO over 2000 miles (3218 km) such as has been recorded in Europe. The greater number of reported 2-meter openings "caught" in recent years is probably due to better monitoring and improved equipment as well as the new common 2-meter calling frequency, 144.2 MHz. The frequency limits for sporadic E are believed to extend above 220 MHz and W1JR and WØVB have both been heard on this frequency by other stations but in each case equipment problems at one end of the path prevented

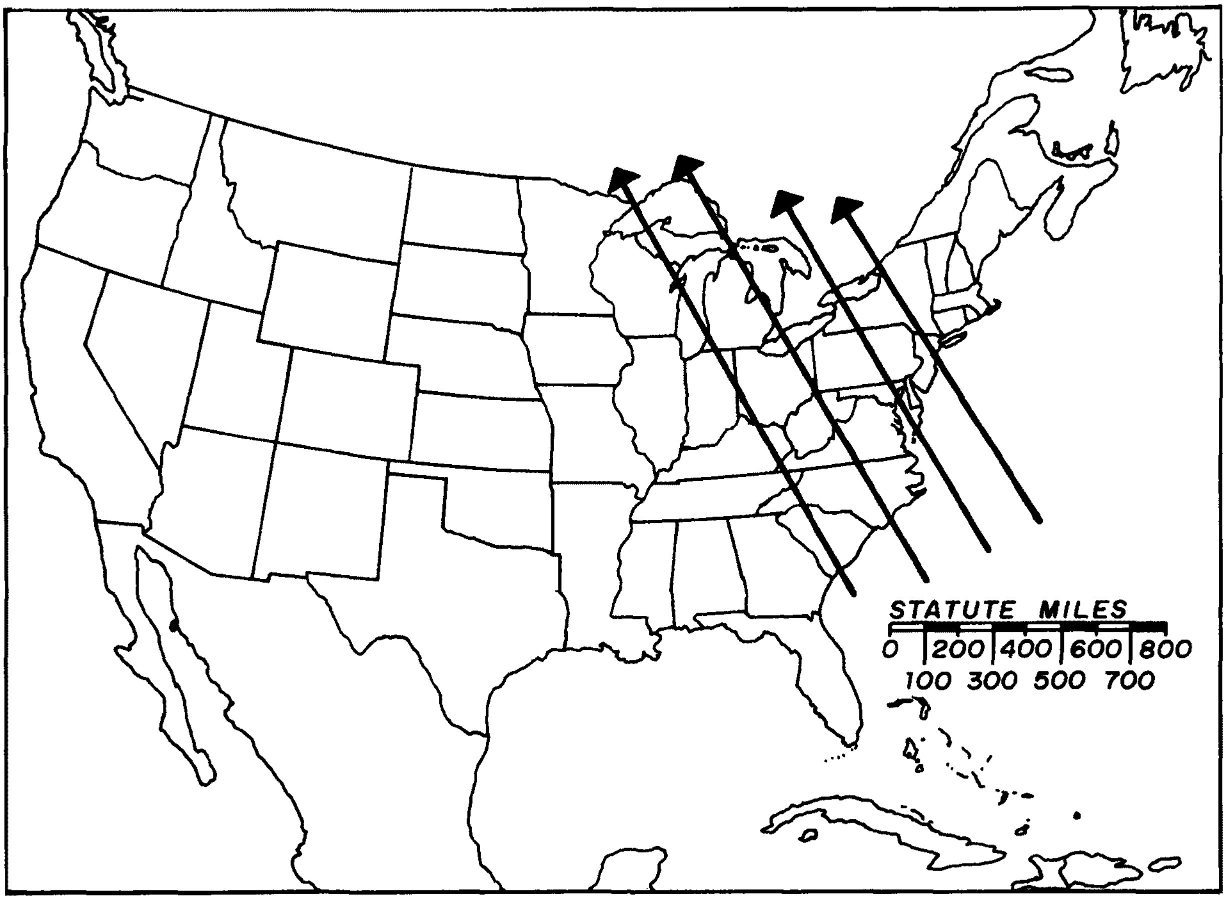

fig. 1. Clouds producing sporadic-E propagation generally travel from southeast to northwest at approximately 180 miles (290 km) per hour, moving in a relatively straight line.

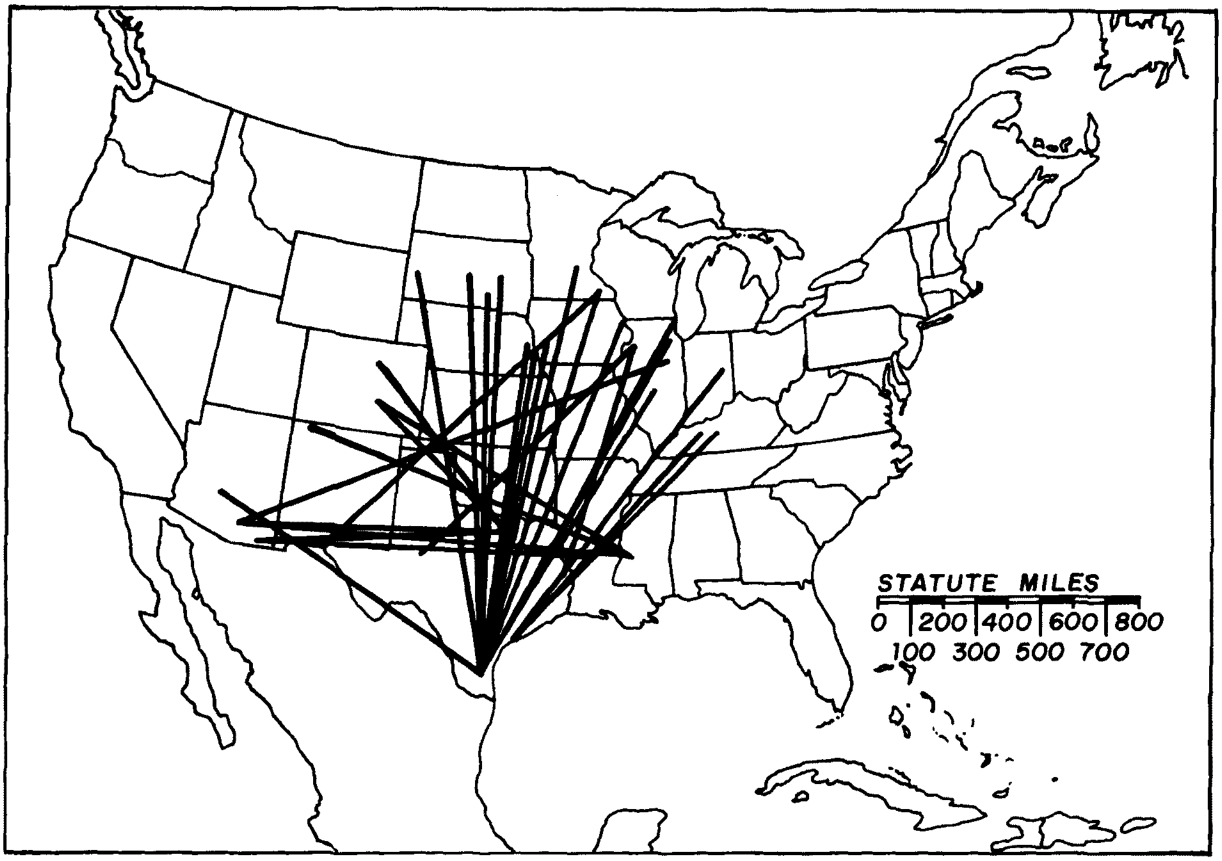

fig. 2. A typical 2-meter sporadic-E opening shows paths between stations in QSO on February 3, 1977, on 144 MHz (0000-0229 UTC).

completed QSOs! (I've noticed, by the way, that all the good east-to-west 2-meter sporadic E openings I've heard from the New England area in the past eight years have occurred within two weeks on either side of July 20th. Any comments?)

The best way to "catch" sporadic E openings is to listen on 10 meters for short skip conditions. When the 10-meter path gets down to 300 to 500 miles (483 to 805 km), the 6-meter band is probably ready to pop or is already open. Similarly, when the path on 6 meters gets very short (perhaps 400 to 500 miles or 645 to 805 km) and intense, try putting out a call on the 2-meter calling frequency. Remember that the reflection from the cloud will usually be different on 2 meters than on 6 meters. The distance will generally be much longer on the higher frequency, so judge accordingly if you're making schedules. Backscatter, especially from a local, is an indication that there are lots of clouds and a big opening is about to begin. You can't have a QSO without making some noise to alert others of an opening. W4WD has noticed that whenever WWV forecasts a "strat-warm," a sporadic E opening usually occurs within the following 12 to 24 hours.

TE (transequatorial) scatter

This was truly an Amateur Radio first when long-distance (5000 miles or 8000 km) propagation was established across the equator on 6 meters in August, 1947. It is significant that tests have *verified that* the signals first enter a bell-shaped ionized area midway between the station and the geomagnetic equator and then are ducted to the opposite ionized area without intermediate reflection from the earth.[6]

For best results, stations should be located about 1500 to 2500 miles (2400 to 4000 km) north and south of the geomagnetic equator and signals should cross the geomagnetic equator at close to right angles. However, deviations of up to ± 20 degrees have been known to occur. It is also significant that propagation may not necessarily be present at the same time at lower frequencies. These contacts usually improve during the equinoctal periods, especially when solar activity is high, as occurs near the peaks of the solar cycles. Openings may start in the late afternoon with clear sounding signals but early evening (at the midpoint on the path) signals often have a flutter with up to a 15 Hz rate. Signals may last until midnight, long

after F2 has disappeared. The frequency usable is typically 1.5 times the MUF maximum that occurred earlier in the same day.

Unfortunately for stations in the continental United States, the geomagnetic equator runs about 15 degrees south of the true equator in this hemisphere. Consequently only stations in very southerly latitudes of the continental United States are optimally located. Stations in South Africa to the Mediterranean as well as Japan to Australia are in more favorable locations for this mode of radio propagation.

equatorial FAI (field aligned irregularities)

There was always the hope that someday there would be propagation as high as 2 meters across the equator. Then on October 29, 1977, it happened, when YV5ZZ/6 in Venezuela made a 2-meter contact with LU1DAU (in Argentina) at a distance of about 3135 miles (5045 km).[7] After this initial success, openings kept occurring and distances kept improving, and by February, 1979, contacts were established between Puerto Rico and Southern Argentina, Australia and Japan as well as between Rhodesia and Greece.

Often the signals had an auroral sound (more on this later). Stations with elevation control noticed that signals were optimized at 5 to 8 degrees elevation. And while 70 cm (432 MHz) signals could be heard no QSOs were made on that frequency. Finally, in 1983, a successful 220 MHz QSO was completed between KP4OER and LU7DJZ (see **table** 1). Some Amateurs speculated on the existence of a new propagation mode,[7] possibly one using a scatter mechanism and FAI from ionized bubbles. Others suggested ducts. At this time no one agrees except to say that more time and study is required.

We now know that propagation between stations equally spaced north and south of the geomagnetic equator can occur up to 70 cm. The best times and seasons for these contacts seem to be those most frequently observed for TE, as explained above. There is much more to be done in this important area — another Amateur Radio *first*.

ionospheric scatter

This mode of propagation,[9] a form of forward scatter, is believed to be due to scattering in the lower D region (30 to 55 miles or 50 to 90 km) of the ionosphere. Typical distances are 800 to 1300 miles (1300 to 2100 km). Signals are usually continuous, but weak and wavery with a broad peak around midday at the midpath and during the summer months. Signal attenuations are approximately 90 dB greater than free-space loss at 6 meters and 115 dB at 2 meters. Hence this mode of propagation is definitely within the capability of two well equipped stations with 2-meter EME capability and deserves more attention.

aurora

Aurora, caused by magnetic disturbances on the sun,[10] is called *aurora borealis* (or "northern lights") in the northern hemisphere and *aurora australis* in the southern hemisphere. Aurora borealis are most common around the equinoxes in March and September and usually begin in the late afternoon or early evening, though they can occur at any time of the year and are sometimes active in the early afternoon as well as all through the night and even into morning! The number of auroral occurrences tends to peak at the beginning and at the end of a solar cycle peak (e.g. 1978-79 and 1982-83) and decrease significantly when sunspot activity is low. They usually occur 24 to 48 hours after a major solar flare. A WWV "A" index of 30 or greater or a "K" index of 4 or more is a good warning of impending aurora. Weak or "watery" sounding HF signals are an excellent indicator of aurora presence although their presence does not guarantee that the openings will extend to VHF/UHF.

Typical auroral propagation occurs in a region between 47 to 84 miles (75 to 135 km) above the earth. If you were to look down on the earth from directly over the magnetic pole, an aurora would appear to be shaped like a halo or doughnut. (See satellite photograph shown in reference 11.) From a vantage point on the earth, auroras appear to be thin sheets or columns of light that are not quite vertical but tipped at the local dip angle of the magnetic field. Those in the far north are usually green, while those that extend further south may be red. The further south the aurora extends, the greater the DX possibilities; they are, after all, a superb reflector of radio waves.

The auroral reflection occurs when the angle of incidence equals the angle of reflection, but incidence and reflection paths need not lie in the same plane. Propagation via auroral reflection has been verified up to about 3000 MHz, but no known Amateur contacts have been reported above 70 cm (432 MHz). The best reported DX is about 1336 miles (2150 km). High power (several hundred watts or greater) is beneficial, but even 10-watt stations have had confirmed 70 cm QSOs. Occasionally the aurora extends far enough south so that even stations in Florida, Texas, New Mexico, and Arizona can establish auroral contacts.

For communications via auroral propagation, both stations should aim their antennas at the same "hot" spot. The cross-section of reflection decreases with increasing frequency. The angle of perpendicularity with respect to the earth's magnetic field also decreases with frequency. Typical accuracies required are ± 12 degrees at 220 MHz, ± 9.5 degrees at 432 MHz and ± 6 degrees at 780 MHz.[12] Since typical antenna installations have beamwidths that usually decrease in both the horizontal and vertical planes with

increasing frequency, the path becomes more difficult and would definitely be improved above 220 MHz if both stations could also change antenna elevation. This is particularly true when trying to contact a station that is either several degrees of latitude north or south of you. Hence, higher power is definitely preferable to increased antenna gain.

Auroral signals usually sound like a "buzz" or white noise because of the Doppler spreading caused by the rapidly changing ionization. CW is preferred, but SSB is usable, though hard to understand and lacking high pitched sounds, if you are operating on the lower VHF bands and the aurora is strong. I've noticed that if a CW signal is narrow (1 to 2 kHz) on 2 meters, there is usually sufficient ionization to permit 220 MHz and maybe even 70 cm contacts. I've also noticed that the MUF usually peaks shortly after the onset of the aurora, perhaps in 15 to 30 minutes, and then slowly decreases as time passes. *Hence it is advisable to take advantage of the higher VHF/UHF bands as soon as the aurora gets going rather than discover later that the MUF has dropped.*

"Watery" or wavery sounding signals on HF (10 to 160 meters) are an excellent indication that aurora is present and can be a tip-off to look for VHF/UHF auroral propagation. Auroras can be easily detected using a TV video carrier monitor on channel 12 (205.25 MHz) or 13 (211.25 MHz), similar to the one I described in last month's column.[13] Just point your beam north and listen for the "buzz" on the carriers. This is also a good system for finding where the aurora is peaking and whether the MUF is high. I've noticed that the "hot" spot may be in a slightly different place at 220 MHz than on 6 or 2 meters (perhaps due to antenna patterns or other variables). This is invaluable information when trying to find 220 MHz and 70 cm signals.

Another observation I've made is that the Doppler shift may vary when you swing your beam back and forth while listening to an auroral signal. I've measured up to ± 1 kHz shift on 2 meters on locals and considerably more on 220 MHz and 70 cm. Consequently it's advisable to tune ± several kHz when listening for answers to a CQ. If you're using a modern transceiver, tune with your RIT so you don't transmit back on a new frequency! Lately I've noticed this happening more often — but doing so can be disasterous on aurora, since you may land on top of another signal and be completely unreadable in the QRM, causing confusion to the operator you've answered, who is no doubt wondering where you went! Likewise, be careful where you zero-beat your own signal if you want an answer.

Finally, look for activity on or near the VHF/UHF calling frequencies. During auroras, this is usually 50.100, 144.100, 220.100 and 432.100 MHz on CW and with 50.110 and 144.2 on SSB. By all means, call CQ. When you hear a reply, try to peak your beam quickly in order to optimize the path, since each station will come in at a slightly different direction depending on latitude and longitude.

artificial auroras

Attempts have been made to generate artificial auroras by using rockets to inject barium clouds into the ionosphere; in recent years, scientists in both the United States and the Soviet Union have created yet another form of artificial radio aurora.[14] This propagation mode is similar to the natural auroras described except in that it is generated by radiating very high amounts (typically 40 Megawatts of ERP) of HF (typically 3 to 10 MHz) CW power vertically into the ionosphere. The scattering center is directly above the transmitter. The principal tests have been conducted in Platteville, Colorado, with a 1-Megawatt transmitter and a large antenna array, and at Arecibo, Puerto Rico, with 100 kilowatts, using the 1000 foot (305 meter) spherical dish. Operations have been successful as high as 430 MHz, and the signals are usually narrower than those generated by natural aurora.

auroral "E_S"

This is a much sought after mode of propagation on 6 meters. According to WAØIQN,[15] ionospheric sounders have verified that sporadic E may be present just below an aurora, but the reasons for this are not known. When particle precipitation becomes strong enough, sufficient electrons collect at the bottom of the field line and form a "puddle" which can spread horizontally. It may then appear like sporadic E after the effects of the aurora wear off and thus yield DX in the northern regions of the United States and Canada. This may explain why some 6 meter auroras sound almost like sporadic E propagation. Auroral E_S usually occur within a few hours after an aurora fades out.

meteor scatter

Meteor scatter is one of the prime modes for DXing on VHF/UHF. Every day 50 to 100 million particles randomly enter the earth's atmosphere, where they burn up in the E region (50 to 150 km high) and leave ionized trails capable of reflecting radio signals. Due to the orbital characteristics of the earth, these random particles tend to increase slowly after midnight, reaching a broad peak at 6 AM and slowly decrease as the morning wears on, reaching a minimum at about 6 PM local time. The best months for random meteors in the continental United States are between June and August, with a minimum occurrence in February.

Even more important to the VHF/UHF DXer are meteor showers. They can occur at any hour of the day, and while they may be short in duration (any-

where from a few hours to a day or two long), they do occur on predictable dates. The better known showers are the Perseids (approximately August 11) and the Geminids (approximately December 11). DX, typically 500 to 1400 miles (805 to 2250 km) is accomplished by reflecting signals off these meteor trails.[16]

Most meteor scatter operation is on 6 and 2 meters, but 220 MHz and occasionally 70 cm (although there have been only about a half dozen reported contacts on this frequency to date) can be used in the higher speed showers. A recent article[13] described how to optimize the use of this mode; readers who wish to know more should refer to that article.

FAI

The authors of reference 7 speculated that FAI similar in nature to that which produces the great TE FAI may be present in the midlatitude regions, and that signals in this mode would probably have an auroral quality. At the same time this article was published, K4GFG was conducting a nightly over-water tropo schedule in an attempt to QSO KP4EOR, who was also involved in the TE FAI experiment. What transpired was a QSO, not by tropo propagation, but by FAI with weak and "watery" signals peaking about 10 to 20 degrees north of the great circle path.[17]

During the following summer (1979), signals appeared over the United States and FAI-type QSOs were made from Florida to Texas and Alabama again, with auroral-type signals peaking north of the great circle path. Occurrences seem to be in the late evening, following sporadic E openings. Typical path loss is in the area of 218 to 230 dB at 2 meters and 248 dB at 220 MHz, meaning that fairly large setups (kW and modest gain antennas) are required to take advantage of this mode. Reference 17 provides charts for pointing antennas and speculates that this mode may be possible as high as 70 cm, although no known QSOs have taken place above 2 meters as of this writing. It is a fascinating possibility for VHFers and a nice way to work real DX when the band would otherwise be quiet!

EME

Earth-Moon-Earth, or "moonbounce" is one of my favorite subjects because it's *the* ultimate DX for the Radio Amateur. When you bounce your signal off the moon and listen for the returning signal, the path is up to 500,000 miles (804,500 km) in length and it takes the signal over 2.5 seconds to traverse the distance! Hence EME allows you to monitor your own signal after it is sent — and if you hear your own echoes, you know that your signal got where it was supposed to go, not like the dead-band syndrome on HF. You also know your equipment is operating successfully; by listening to the strength of the echo you can also tell how well it's working and whether your most recent improvement actually works.

While EME is not for the faint-hearted, it can be achieved by the typical VHF/UHFer. The first EME stations were highly experimental and results were marginal at best. However, with the advances in the state-of-the-art (such as good low-noise transistors and GaAs FETs, high performance Yagis and parabolic dishes as well as efficient amplifiers), EME really took off in the late 1960s. Nowadays, when conditions are good on 2 meters, you have the capability of hearing your own echoes; WAS as well as WAC are possible with 500 watts of output power and four 3.2 wavelength NBS-type Yagis. Even a single-Yagi station with a good low-noise antenna-mounted preamplifier can hear the larger EME 2-meter and 70 cm stations off the moon when conditions are good. Over 50 DXCC countries are now or have been active on EME in the last few years and EME expeditions are no longer uncommon.

Typical path losses for EME range from 251 dB on 2 meters to 276 dB on 13 cm (2300 MHz), the highest frequency band on which Amateur contacts have been claimed.[18] A few EME QSOs have been made on 6 meters, but the antenna systems were *very* large, and noise was a limiting factor. EME is now commonplace on 2 meters, 220 MHz, 70 cm, and 23 cm, where contacts are made almost daily and even SSB QSOs are not uncommon. Typical EME operating and calling frequencies are listed in **table 2**.

table 2. Typical EME operating and calling frequencies.

band	frequencies used	calling frequencies
2 meters	144.000-144.110	144.003-144.010
220 MHz	220.0-220.080	220.020
70 cm	432.000-432.070	432.010
23 cm	1296.000-1296.060	1296.010
13 cm	2304-2304.1 and 2320-2320.2	2304.050 and 2320.150*

*Most Europeans can no longer operate below 2320 MHz in the 13 cm band. The USA Amateurs may soon lose the frequencies between 2310-2390 MHz. Therefore, cross band operation may be required.

13 cm is now being tried, and DFØEME has reported several QSOs using a 9-meter dish. The interesting thing about 23 and 13 cm is that the power levels and antenna systems are small in comparison to 2 meters and 220 MHz. For example, some stations on 23 cm have been making QSOs rather routinely with dishes as small as 8 to 13 feet (2.5 to 4 meters) in diameter and with power as low as 100 watts at the feed. They're making good use of the new low-noise GaAs FET preamplifiers mounted right at the feed and of the

very low sky noise above 400 MHz. All things being equal, path loss and antenna gain increase directly with frequency. However, since the antenna gain is added twice (once on receive, and once on transmit), the overall signal-to-noise ratio increases with increasing frequency. Another advantage of EME is that you don't have to wait for a band opening!

The EIMAC EME notes[19] are a great help for the newcomer to EME. Reference 20 describes the requirements for 70 cm EME as well as helpful hints to improve your station and equipment both on EME and terrestrial. (A later column will describe 220 MHz requirements. See you off the moon!)

tropospheric scatter

This mode of propagation uses the reflections off dust particles, clouds and the refractive index variations that occur in the troposphere (1000-50,000 ft. or 305-15,250 meters above sea level) to provide reliable VHF/UHF propagation up to 1000 miles (1609 km) between well equipped stations. [21,22] For several years starting in 1953 Chisolm et al[22] conducted interesting tropo scatter tests over 98-830 mile (158-1335 Km) paths from Massachusetts to North Carolina. They proved that with adequate equipment, 70 cm signals could always get through with path loss varying from 190 dB at 98 miles (158 km) to 258 dB at 618 miles (994 km) and 300 dB at 830 miles 1335 km).

Both stations should aim their antennas at each other on the great circle path and their signals will scatter from the common volume of the atmosphere somewhere near the mid point in the path. For best long haul DX, a low angle of radiation is essential making it desirable to have your antenna fairly high (10 wavelengths minimum but greater than 20 wavelengths is unnecessary), in the clear, and a good low-angle horizon. Antenna gain is not fully realized since the power is not from a point source but from a volume in the atmosphere.

Typical path attenuations for the long DX (greater than 300 miles or 483 Km) are such that an EME type setup is desirable. However, many EME stations are now using their EME arrays with mast mounted preamplifiers so DX is no longer a problem on the VHF/UHF frequencies. Recent work in the U.K. on 3 cm (10 GHz) shows that troposcatter is a very reliable mode.[21] They predict that a 440 Km range is possible using 1 watt and 4 foot (1.22 meter) dishes by each station. This propagation mode deserves more attention especially on 70 cm and above!

tropospheric bending

Tropospheric communication utilizes weather-related changes in the atmosphere to propagate VHF/UHF signals over much greater than line-of-sight distances. Under normal weather conditions, the temperature of the atmosphere decreases in a more or less linear fashion with increasing altitude. However, if the temperature should abruptly increase with increasing altitude, the conditions are right for long haul DX. This usually happens between 3000 and 6000 feet (915 to 1930 meters) above local terrain and causes the refractive index to significantly increase beyond 1.33 as started earlier in the line-of-sight discussion. If you are at an elevated QTH, you could get above the temperature inversion and be unable to propagate through this layer!

Best tips are to look for increased cloudiness following a period of fair and calm weather. Best periods occur after a long (two or more days) high steady barometer (greater than 30.3 inches or 1025 millibars) starts to drop due to the approach of a slow moving low pressure area from a few hundred miles away, especially when moist air masses are approaching such as from the Gulf of Mexico. Signals are usually strongest in the late evening and especially in the early morning around sunrise. For advanced warning watch weather maps especially on TV. Remember that weather maps in the newspapers may be 24-48 hours old and hence may be after the fact! The late Ross Hull wrote several interesting articles on this subject.[23]

I have also noticed that the really good overland tropospheric bending generally occurs in the winter months in the states bordering or near the Gulf of Mexico and between mid-August and late November in the central, northern and northeastern continental USA. Openings are also prevalent when there are large inversion layers with uncomfortably high humidity. The eastern USA should look for enhanced north to south DX when a weather condition called a "Bermuda High" is present. The best DX, especially east to west, seems to occur when a hurricane is bearing down on the area. Fortunately or unfortunately, the hurricanes that used to come roaring across the eastern half of the USA have been few in numbers since the early 1970s.

tropo ducting

This form of propagation is quite similar to the tropospheric bending just discussed but is more prevalent on the over water paths. In contrast to the normal tropospheric bending caused by an abrupt temperature change with increasing altitude, this form of propagation usually has two abrupt changes, one very low (maybe only a few meters above the surface) and the other one above it by perhaps 1000 feet (300 meters) or so. The net result is that a sort of waveguide is formed, the thickness of which determines the best frequencies of radio propagation. If you are in the duct, signals will propagate almost unattenuated until the duct deforms. If you are outside, take up another pastime!

Those who have studied ducting of radio signals tell me that at the onset of the duct, the optimum frequency is often quite high, typically 1500 MHz, and may slowly decrease as the duct stabilizes often going up again as the duct starts to disappear. During the great tropospheric ducting conditions between Hawaii and California in July, 1973, the openings were first heard on 70 cm and then slowly shifted to 2 meters. Two days later even 6-meter signals (viz. KH6IJ) were making the grade. It is noteworthy that this duct took place while two tropical (hurricane type) storms were traversing the path between Baja, California, and Hawaii just to the south of the path.

There are other pecularities about tropo ducting. If you are above or below the duct (as in tropospheric bending), you will not be able to couple sufficiently into the duct to take advantage of it. Seldom does an over-water duct extend very far inland. Instead, the duct may abruptly become elevated as it passes over land. Hence, if you are over 10 to 40 miles (16 to 64 km) inland, the only effective way to use the duct will be if you are elevated or have a reflection assist (more on this later).

Let me elaborate on the later point. Researchers in the United States Navy and others have studied the California to Hawaii ducting with aircraft and often found that the duct was elevated 1 to 2 miles (1.6 to 3.2 km) as it approached Hawaii. The Amateurs active in Hawaii frequently drove up and down the side on Mauna Loa to optimize this path. After several years of observing this path, they installed the 24-hour-a-day warning beacons well up the side of Mauna Loa to take advantage of this phenomenon. Unfortunately, if an opening takes place (as seen by monitoring the beacon from California), there has to be an able body to drive up to the sight as soon as possible to catch the opening, and this hasn't always been possible!

The farthest inland ducting I know of is the KD6R to KH6IAA QSO. Ironically, both stations were elevated. KD6R was about 38 miles (61 km) inland, but at an elevation of almost 6000 feet (1828 meters) on the side of Mount Palomar. Likewise, the Bermuda-to-USA path where coastal stations can often access the duct but inland stations can't except K2RIW who is about 10 miles (16 km) inland but elevated on one of the highest spots (400 feet or 121 meters) on Long Island. K1PXE, at sea level in Connecticut, has never heard Bermuda, perhaps because the signals have to first traverse Long Island and are thereby elevated beyond reasonable altitudes.

The best ducts observed and used for long DX are the "Great Australian Bight" off southern Australia, the Gulf of Mexico, the Mediterranean Sea, the North Sea, Bermuda to the United States, the Atlantic Ocean between the United Kingdom and the Canary Islands, and of course the California-to-Hawaii path. In fact, rumor has it that a 5 GHz microwave signal from the Philippines was intercepted in Southern California in the mid 1970s but that the news was suppressed because no one would believe the story! There is still much to be learned about tropospheric ducting and perhaps someday the United States-to-Europe path will finally be conquered using this mode.

super refraction

With a few exceptions, this mode is really an extension of tropospheric ducting. These ducts are usually very intense, over warm water, close to the surface and probably not as thick as the usual tropospheric ducts. Hence they are primarily good at 23 cm and above. The British were the first ones I know to really exploit this mode. In their somewhat casual manner (as we see it!) they would go to the beach for a Sunday outing with a pair of "GunnPlexers" or similar 3 cm gear. During the day they would occasionally go down to the beach and try to communicate with a station on another beach over an all-water path. By trial and error they would often find an optimum height and time for a QSO.

Another interesting story I've heard also comes from the UK. The English Channel is an ideal spot to try UHF between England and the European continent. As the story goes, an English station was in 3 cm communications across the channel with a French station, when communications abruptly ceased and then returned just as quickly a few moments later. Upon observation of the path with binoculars, it was discovered that a larger ship had apparently passed between the stations and momentarily broken up the duct!

Not to be outdone, the Italians and Yugoslavians used the Adriatic and Mediterranean Seas. They had many successes, the greatest DX being the 3 cm QSO between I0SNY/EA9 and I0YLI/IT9 (see **table** 1). This was an incredible DX record of 1034 miles (1663 km), using only 50 milliwatts of power and 1 meter dishes.[24] Once again, there is much to be learned about super refraction, but what is clear is that UHF/SHF ducts do appear on over-water paths during warm weather, especially during the summer months.

lightning scatter

The first known QSO via this mode took place on 70 cm between W0DRL (KS) and W5RCI (MS) on September 16, 1968, over a 449 mile (722 km) path. They noticed that signals peaked 15 to 18 degrees off the great circle path at an area which turned out to be a storm cell over Texas. Enhancements were up to 40 dB, some with durations of 25 seconds, with extremely rapid QSB and lots of Doppler shift. Since then, other observers have noticed the same phenomenon. This is a mode worth looking into, especially

when severe lightning storms are in progress. I have observed this same type of propagation on 70 cm on somewhat shorter paths between New Jersey and Massachusetts when severe lightning was over the central Connecticut area. Since most VHFers tend to shut off their rigs during stormy weather, it may be well to rethink our operating habits — but only if the storm *isn't* nearby!

aircraft scatter

Few VHF/UHFers give this mode a thought or even are aware of its capabilities. While in California, I discovered (undoubtedly as others did), that aircraft make fine reflectors for radio signals, especially on 70 cm and above. Since commercial aircraft often fly as high as 40,000 feet (12,192 meters), they can be used for propagation via line-of-sight out to 500 miles (805 km) reflections on a regular basis. The best paths are over the central USA, since many aircraft are available and usually fly at high altitudes. Coastal QTHs are not as favored because these aircraft usually fly low or are in takeoff or landing patterns. But don't rule them out; the California coast and the eastern seaboard of the United States have lots of traffic though these areas aren't quite as good for the long DX via aircraft scatter.

All you have to do to work via aircraft scatter is to be alert, know when the flights are in between and set up a schedule! Don't use long transmissions: at the longer DX, the mutual location for best scatter may last only a minute or less! Beware also of the Doppler shift which can be as much as 100-300 Hz on 70 cm and up to 1.5 kHz on 2304. When I QSO'd Harley, WA6HXW, on 2304 MHz from W6FZJ, a path of 310 miles (500 km), I copied Harley on two different frequencies separated by almost 1.5 kHz, one via tropo and the other via aircraft. At times the aircraft-reflected signal was up to 5 degrees off the great circle path!

table 3. Typical selected aircraft reflection sizes and figures of merit (original source of data is unknown but was sent to me via K9MHR and WB0YSG). Intermediate sized aircraft can be easily estimated.

type of aircraft	reflecting area (square meters)	relative reflection (dB)
Cessna 336 Skymaster	1.3	1
Lear Jet	2.0	3
McDonald Douglas DC 9	8-10	9-10
Douglas DC 3	12.6	11
Boeing 707	16.0	12
McDonald Douglas DC 8	20.0	13
Boeing 747	63.0	18

Some relative data on aircraft sizes and figures of merit are listed in **table 3**. This data, which may be of use to those who want to make radar path loss calculations, show why larger aircraft are preferred. I have also noticed that aircraft scatter can often assist with another mode for a sort of link-up. For instance, on both the Bermuda to USA and the California to Hawaii paths, my 70 cm signals have been coupled into the duct and copied, probably via aircraft scatter at my end, but I was unable to complete the QSO due to insufficient access time before signals disappeared. This is an entertaining mode that deserves more attention since it is available 365 days a year as long as you have a reasonably good station. The possibilities for long DX on 3 cm and above are fascinating.

knife-edge diffraction

This mode of propagation, used for many years for Gigahertz microwave communications by commercial stations, is well documented. It is based on the theory that if a sharp peak or hill lies between two stations, signals can be diffracted over them. Losses of only 10 to 20 dB over the typical free space path loss are possible.[2] The sharper the peak, the better. Lack of foliage or reflecting objects is preferred at the peak; 3 cm signals do not go through tree leaves! Although this is a specialized propagation mode and is primarily for moderate DX, it is nevertheless interesting, especially to the UHFer who is in a low lying area or surrounded by hills or mountains.

rain scatter

This mode of propagation was discovered in 1978 by G3JVL and G3YGF/A when operating on the 3 cm band over a 110 km path during a rain squall.[25] They found that if two stations both aimed their antennas at the storm center, they could communicate at distances greater than 100 km. They were unable to reproduce this mode at lower Amateur frequencies. The signals are highly distorted and have an auroral quality with some noticeable Doppler. Signals tend to peak very broadly in azimuth and elevating the antenna can improve signals if the storm is nearby. 10-20 dB enhancements are typical. Other British Amateurs also participated and verified this phenomenon, even with low power stations. This mode is especially interesting when one or both of the stations have obstructed views or a poor VHF/UHF QTH since the scattering center may be well above the horizon at one or both ends of the path. Hopefully this mode will soon be more fully utilized here in the USA.

conclusion

VHF/UHF radio propagation is a fascinating subject to which every Amateur can contribute. All it takes is time, patience, good notes, and of course some reasonably good equipment. The records listed in **table 1** are good goals to pursue.

To review, VHF/UHF radio propagation is basically divided into four categories:

1. *Natural* (line-of-sight, tropo/ionospheric scatter),
2. *Weather-related* (tropospheric bending, ducting, sporadic E, precipitation scatter),
3. *Celestial* (EME, meteor scatter, F2, aurora), and
4. *Manmade* (satellite, aircraft scatter, artificial aurora).

A basic knowledge of all these types of radio propagation will show that the VHF/UHF bands are hardly a quiet and uninteresting place!

To predict radio propagation conditions we can use several sources of data such as: HF/VHF or satellite nets, propagation beacons, TV video carriers monitoring or the warning sounds on HF signals as well as alerts such as weather maps/reports and WWV forecasts at 18 minutes after each hour.

table 4. These are the typical United States VHF/UHF calling frequencies. It is common courtesy to QSY up or down after establishing contact to allow other stations to use the calling frequency.

band	frequency	mode
6 meters	50.110	SSB
2 meters	144.100	CW
2 meters	144.200	SSB
220 MHz	220.100	CW/SSB
70 cm	432.100	CW/SSB
23 cm	1296.100	CW/SSB
13 cm	2304.100	CW/SSB

Calling frequencies listed in table 4 are also helpful. Sometimes there just aren't enough signals to go around and congregating on special frequencies can increase the probability of observing good conditions. However, you must put out an occasional call because silence doesn't convey much activity! Above all, when you do establish contact on a calling frequency, move off so that someone else can use it. Activity nights are also helpful. In the northeast, we concentrate 2 meter activity on Monday evenings, 220 MHz on Tuesdays, 70 cm on Wednesdays, and 23 cm on Thursdays.

acknowledgements

I want to thank all those who helped contribute information for this article, and in particular G3WDG, VE1YX, W4WD, WA4MVI, WB5LUA, W6ABN, and K6FV — it takes lots of cooperation to gather so much material. I hope that some of this information will be new or helpful to you and that you too will be able to help in the future.

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VHF/UHF coming events

July 3: EME perigee

July 20 (± 2 weeks): look for 2-meter E openings

July 27: (21:15 UTC) peak of Delta Aquarids Meteor Shower

July 27-29: Central States VHF Society Conference, Cedar Rapids, Iowa. (Contact WØOHU for further information.)

July 31: EME perigee

If you have any VHF/UHF events or contests planned, let me know at least 2 to 3 months in advance so I can let others know through this column. — **W1JR**

ham radio

cooling semiconductors part 1: designing and using heatsinks

Can't stand the heat? dissipate it!

What effect does heat have on semiconductors? How can that heat be dissipated? And what are the implications, for designers of electronic equipment, of the answers to these questions? Part one of this series will cover the design and use of heatsinks; part two will address other means, both novel and conventional, of monitoring and cooling semiconductors *without* the use of semiconductors — for example, by using a fan with flapping blades that operates on the piezoelectric crystal principle.

It is crucial for transistors, ICs, or thyristors (triacs and SCRs) to remain within their safe operating junction temperature ranges. Semiconductor junctions can withstand temperatures between 85 degrees C and 200 degrees C. Germanium withstands from 85 degrees C to 100 degrees C maximum temperature, while silicon withstands from 150 degrees C to 200 degrees C maximum temperature. However, semiconductors operated at these high temperatures show reduced performance. For example, low-input bias op amps require twice as much bias current for every 10-degree C rise in junction temperatures. (Note the typical power derating curve in **fig. 1** for a power transistor.)

Both junctions in a transistor give off heat; however, since the collector-base junction is reverse biased, it has higher resistance and produces more heat. In fact, the collector-base junction produces so much more heat than the transistor's forward-biased emitter-base junction that the heat generated by the later junction is usually ignored. As a result, the power dissipation equation is simplified and can be defined as follows:

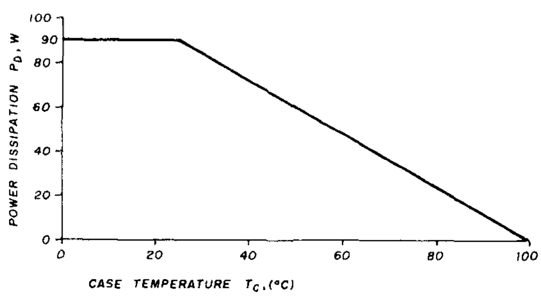

fig. 1. A dissipation derating curve for a typical power transistor housed in a TO-3 case.

$$P_D = I_C \times V_{CE} \tag{1}$$

where P_D = power dissipation
I_C = collector current
V_{CE} = voltage from collector to emitter

electrical-thermal analogies

Analogies can be helpful in visualizing abstract phenomena. For example, high school science teachers will often compare electricity to water flowing from a tower. In this comparison, voltage is defined as the "push" or potential (height) of the water within this tower. Current is equated with flow of water, and resistance is compared to the action of the tower's faucet, which impedes or constricts flow.

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Table 1. Unit symbols in electrical and thermal models.

electrical			heat		
quantity	unit	symbol	quantity	unit	symbol
voltage	V	V	temperature	°C	T
current	A	I	heat flow	W	P_D
resistance	Ω	R	thermal resistance	°C/W	θ

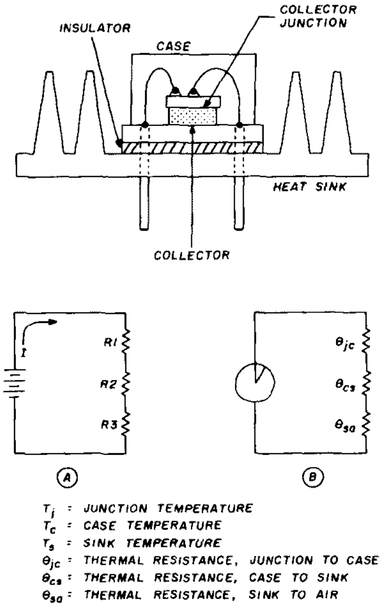


fig. 2. An electrical-thermal analogy.

An analogy can also be drawn between heat transfer and electrical parameters (**fig. 2**). The flow (transfer of heat) corresponds to current. The heat differential between the hottest object (the transistor's junction) and the coolest object (the free-standing ambient air) is seen as voltage, and each thermal resistance — from the resistor's junction to its case, to the heatsink, and eventually to the free-standing ambient air — is compared to electrical resistances in series. Therefore, thermal resistances are added together to determine the total thermal resistance.

In **fig. 2**, the dual subscript on a thermal resistance, such as θ_{SA}, means resistance of the sink (S)-to-ambient air (A). **Table 1** illustrates the unit symbols in both the *electrical and thermal models.*

The basic thermal relationship is:

$$T = P_D\theta \qquad (2)$$

where T = temperature rise
P_D = power dissipation
θ = thermal resistance

It becomes obvious from this equation that thermal resistance θ must be minimized. *This is what a heatsink does.*

how heatsinks work

Power transistors and voltage regulators with 1A or greater ratings are usually installed in large metal cases so that they can radiate more heat. Because the case size may still be inadequate for the amount of power to be dissipated, a heatsink or similar metal fitting is added to the case to increase thermal mass, lower thermal resistance, and provide additional surface area to allow heat to radiate from the surface more effectively.

The purpose of a heatsink is to transfer heat away from the semiconductor, causing it to flow from one medium to another by means of one or more of the three heat transfer methods: conduction, convection, and radiation.

Conduction is the most effective means of moving heat from the semiconductor junction to the case and from the case to the heatsink. *Convection* — which is merely surface conduction from a surface to a fluid or to air — is the most effective means of moving heat from case or heatsink to ambient air. *Radiation*, the most effective means of transferring heat from radiating heatsink fins, depends on the existence of a difference in temperature between radiating and adjacent objects or space. *Emissivity* is a term used to describe this effect; **table 2** illustrates some typical emissivities of various common surfaces. (The greater the emissivity, the better the ability to radiate heat.) Note that black oil-painted heatsinks radiate most effectively; this is why commercially available heatsinks (**fig. 3**) are almost always painted black.

heatsink materials

The two most common materials for heatsinks are copper and aluminum. Copper has a thermal conductivity four times that of aluminum, but is much more expensive. Aluminum, therefore, is the most commonly used heatsink material. In applications in which weight is a concern, magnesium can be used. If insulation is necessary, beryllium oxide (BeO), an excellent electrical insulator, can be used.

Very often, power transistors such as the ones packaged in TO-3 or TO-66 cases ("TO" stands for "transistor outline"; the number identifies the shape and size of the case) have their collectors soldered directly to the cases. Unless the collector is to be electrically grounded, the case must be isolated from the chassis.

Three other materials (used in electrical insulating washers) are *also* effective heat conductors: mica,

anodized aluminum, and beryllium oxide. Mica is the most common. Although anodized aluminum is fairly widely used, an electrical short will result if its surface is scratched. Beryllium oxide performs better than mica or anodized aluminum, but both its powdered form and its fumes are toxic.

custom made heatsinks

For the homebrew enthusiast in electronics who wants to make his or her own heatsinks, the nomographs of metal types and shapes and fin effectiveness, shown in **figs. 4** and **5**, will aid in the selection of appropriate materials and forms. Perhaps you have access to some surplus extruded copper or aluminum, or just want to attach a single flat square or rectangular piece of metal to a voltage regulator or power transistor and wonder whether it will do the job. A good question!

Before making any calculations, identify the semiconductor's case. If its case is a TO-220 to TO-202 metal tab, look at the accompanying data sheet to see whether the metal tab is the collector on a transistor or the output pin on a voltage regulator, or if it is the emitter (which might be grounded) or the ground pin on a voltage regulator. If the metal surface of the semiconductor's case had *any* voltage indicated on it, paint the heatsink — or chunk of metal you are about to fashion into a heatsink — so that it will be electrically insulated. A mica insulating wafer or washer may also be used, but will add a small thermal resistance.

Will a heatsink be necessary? Is the one you have in mind adequate? Use this typical heatsink calculation to be sure. A generalized equation of temperature differential (rise), power dissipation, and thermal resistance is:

$$P_{D(MAX)} = \frac{T_{J(MAX)} - T_{A(MAX)}}{\theta_{JA}} \qquad (3)$$

where $T_{J(MAX)}$ = maximum junction temperature in degrees centigrade
$T_{A(MAX)}$ = maximum ambient air temperature in degrees centigrade
θ_{JA} = junction-to-air thermal resistance

This equation states that the maximum power dissipated is equal to the maximum junction temperature of the transistor or regulator minus the maximum ambient or outside free-standing air temperature divided by the thermal resistances from the junction (J) to the ambient air (A). These thermal resistances, analogous to electrical resistances in series, are added together. Stated mathematically, $\theta_{JA} = \theta_{JC} + \theta_{CS} + \theta_{SA}$; the thermal resistance from junction to air (θ_{JA}) equals the sum of the thermal resistances:

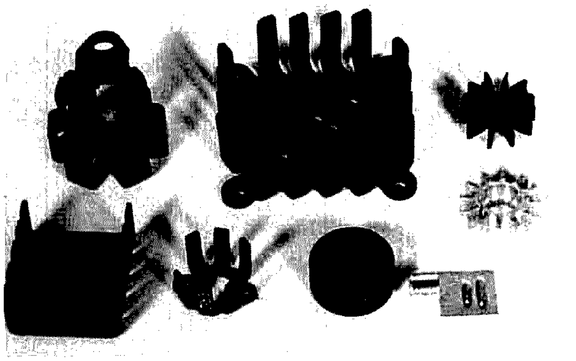


fig. 3. Commerically-available heatsinks vary in form, but are almost always painted black.

junction-to-case (JC), case-to-heatsink (CS) and heatsink-to-air (SA). Quite simply, if the $P_{D(MAX)}$ of the preceding equation is exceeded, then a heatsink is required. Typically, a TO-3 case will withstand 2.8 watts, and a TO-220 power tab with a flat rectangular body and three parallel leads coming out of the end, separated by 0.1 inch each, will withstand 1.8 watts, while the very small TO-39 and TO-92 packages will handle only about 0.66 watts.

Assume the following in our heatsink calculations:
maximum ambient temperature: 60 degrees C
maximum junction temperature: 125 degrees C
maximum output current: 800 mA
maximum input-output differential: 10 volts

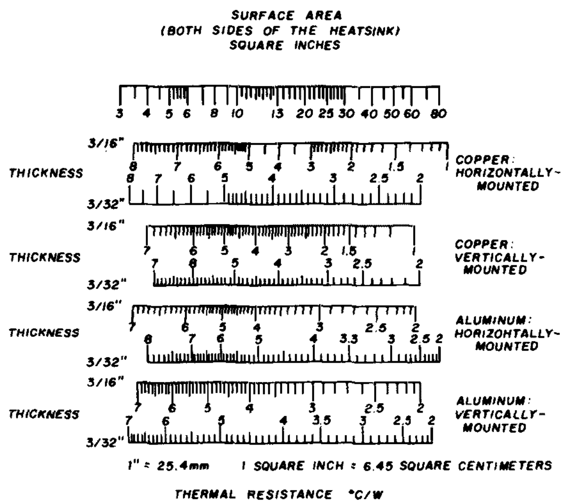
If we use a 7800 series voltage regulator with a TO-220 case, $\theta_{JC} = 5°C/W$ (this is obtained from the data sheet). The θ_{CS} and θ_{SA} must be solved for by:

$$\theta_{JA} = \theta_{JC} + \theta_{CS} + \theta_{SA} = \frac{T_J - T_A}{P_D}$$

$$= \frac{125°C - 60°C}{0.8\ A \times 10\ V} = \frac{65°C}{8\ watts} = 8.13°C/W \qquad (4)$$

where θ_{JC} = junction-to-case thermal resistance
θ_{CS} = case-to-heatsink thermal resistance
θ_{SA} = heatsink-to-air thermal resistance
T_J = junction temperture
T_A = ambient air temperature

Let us solve the far right side of the equation first, $T_J - T_A = 65°C$. The denominator is the 800 mA output current times 10 volts, the input-output voltage differential. The three thermal resistances, then, sum to 8.13°C/W. From the data sheet we already know that the junction-to-case thermal resistance is 5°C/W, so the other two remaining resistances must add up to 3.13°C/W. With the use of silicon thermal grease (**table 4**), we can approximate with some assurance from **table 1** that the case-to-sink resistance is



NOTE: TO DETERMINE EITHER AREA REQUIRED OR THERMAL RESISTANCE OF A GIVEN AREA, DRAW A VERTICAL LINE BETWEEN THE TOP (OR AREA) SCALE DOWN TO THE MATERIAL OF INTEREST. TO DETERMINE THE REQUIRED SURFACE AREA FOR A GIVEN MATERIAL WITH A KNOWN THERMAL RESISTANCE, DRAW A VERTICAL LINE FROM THE APPROPRIATE SCALE UP TO THE SURFACE AREA SCALE.

fig. 4. Thermal effectiveness of various types and thickness of metals.

Table 2. Emissivity of selected common heatsink surfaces.

Surface	emissivity (E)
polished aluminum	0.05
polished copper	0.07
rolled sheet steel	0.66
oxidized copper	0.70
black anodized aluminum	0.7–0.9
black air drying enamel	0.85–0.91
dark varnish	0.89–0.9
black oil paint	0.92–0.96

0.13°C/W, so this leaves the $\theta_{SA} = 3°CW$. Refer again to **fig. 4** and note that a 22 square-inch piece of aluminum 3/16-inch thick with vertically bent ends will meet this requirement. The note at the bottom of the figure explains how to use the nomograph.

Calculating the effectiveness of a heatsink with fins is much more involved than calculating the effectiveness of a heatsink of square or rectangular metal as was just done. The heatsink-to-ambient air thermal resistance is expressed as:

$$\theta_{SA} = \frac{1}{2H^2\eta(h_c + h_r)} \;°C/W \qquad (5)$$

where H = vertical height of fins in inches
η = fin effectiveness factor
h_c = convection heat transfer coefficient
h_r = radiation heat transfer coefficient

$$h_c = 2.21 \times 10^{-3}\left(\frac{T_S - T_A}{H}\right)^{1/4} W/in^2°C \qquad (6)$$

$$h_r = 1.47 \times 10^{-10} E\left(\frac{T_S + T_A}{2} + 273\right)^3 W/in^2\,°C \qquad (7)$$

where T_S = temperature (°C) of heatsink at regulator mounting
T_A = temperature (°C) of ambient air
E = surface emissivity (refer to **table 2**)
H = vertical height of fins in inches

do it yourself

Using **fig. 5** and the formulas shown, let's design a finned heatsink for a 5-volt regulator. Assume that we are using a *1/16-inch thick piece of black anodized* aluminum and want to have symmetrical square heatsink fins. Our desired heatsink will have a thermal resistance of 4°C/W. An LM340T-5 is used under the following realistic conditions:

$$T_J = 125°C \qquad V_{OUT} = 5V$$
$$T_A = 60°C \qquad I_{OUT} = 800\ mA$$
$$V_{IN} = 15V$$

We find T_S by $I \cdot V \cdot \theta_{JS} = T_J - T_S$. V equals $V_{IN} - V_{OUT}$ or $(15V - 5V) \cdot 800\ mA \cdot 4°C/W = 125°C - T_S$. Solving for T_S we obtain T_S equals 93°C. Next h_c is solved for using a 4-1/4 inch high fin.

$$h_c = 2.21 \times 10^{-3}\left(\frac{93°C - 60°C}{4.25\ in}\right)^{1/4} = 3.7 \times 10^{-3}$$

and h_r is solved for with $E = 0.9$, which is found in **table 2**.

$$h_r = 1.47 \times 10^{-10} \times 0.9 \left(\frac{93°C + 60°C}{2} + 273\right)^3$$
$$= 5.6 \times 10^{-3}$$

Next, add h_c to get $h = 9.3 \times 10^{-3}$. Now we find η from **fig. 5**. Draw the first line on the nomograph in **fig. 5** from h through the fin thickness (remember, we are using 0.0625-inch (1/16-inch) aluminum to find

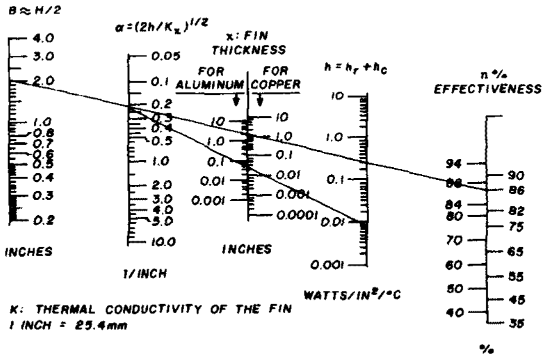


fig. 5. Fin effectiveness nomogram for symmetrical, flat, uniformly thick, vertically mounted finned heatsinks.

table 3. Commercially available heatsinks.

No attempt has been made to provide a complete list of all heatsink manufacturers. This list is only representative.

θ_{SA} Approx[1] (°C/W)	Manufacturer & Type
For TO-202 Packages	
12.5 - 14.2	Staver V4-3-192
13	Staver V5-1
15.1 - 17.2	Staver V4-3-128
19	Thermalloy 6106 Series
20	Staver V6-2
25	Thermalloy 6107 Series
37	IERC PA1-7CB with PVC-1B Clip
40 - 42	Staver F7-3
40 - 43	Staver F7-2
42	IERC PA2-7CB with PVC-1B Clip
42 - 44	Staver F7-1
For TO-220 Packages	
4.2	IERC HP3 Series
5 - 6	IERC HP1 Series
6.4	Staver V3-7-225
6.5 - 7.5	IERC VP Series
8.1	Staver V3-5
8.8	Staver V3-7-96
9.5	Staver V3-3
10	Thermalloy 6032, 6034 Series
12.5 - 14.2	Staver V4-3-192
13	Staver V5-1
15	Thermalloy 6030 Series
15.1 - 17.2	Staver V4-3-128
16	Thermalloy 6106 Series
18	Thermalloy 6107 Series
19	IERC PB Series
20	Staver V6-2
25	IERC PA Series
26	Thermalloy 6025 Series
For TO-92 Packages	
30	Staver F2-7
46	Staver F5-7A, F5-8-1
50	IERC RUR Series
57	Staver F5-7D
65	IERC RU Series
72	Staver F1-7
85	Thermalloy 2224 Series

θ_{SA} Approx[1] (°C/W)	Manufacturer & Type
For TO-5 Packages	
12	Thermalloy 1101, 1103 Series
12 - 16	Wakefield 260-5 Series
15	Staver V3A-5
22	Thermalloy 1116, 1121, 1123 Series
22	Thermalloy 1130, 1131, 1132 Series
24	Staver F5-5C
26 - 30	IERC Thermal Links
27 - 83	Wakefield 200 Series
28	Staver F5-5B
30	Thermalloy 2227 Series
34	Thermalloy 2228 Series
35	IERC Clip Mount Thermal Link
39	Thermalloy 2215 Series
42	Staver F5-5A
45 - 65	Wakefield 296 Series
46	Staver F6-5, F6-5L
50	Thermalloy 2225 Series
50 - 55	IERC Fan Tops
51	Thermalloy 2205 Series
53	Thermalloy 2211 Series
55	Thermalloy 2210 Series
56	Thermalloy 1129 Series
58	Thermalloy 2230, 2235, Series
60	Thermalloy 2226 Series
68	Staver F1-5
72	Thermalloy 1115 Series

θ_{SA} Approx[1] (°C/W)	Manufacturer & Type
For TO-3 Packages	
0.4 (9″ length)	Thermalloy (Extruded) 6590 Series
0.4 - 0.5 (6″ length)	Thermalloy (Extruded) 6660, 6560 Series
0.56 - 3.0	Wakefield 400 Series
0.6 (7.5″ length)	Thermalloy (Extruded) 6470 Series
0.7 - 1.2 (5 -5.5″ length)	Thermalloy (Extruded) 6423, 6443, 6441, 6450 Series
1.0 - 5.4 (3″ length)	Thermalloy (Extruded) 6427, 6500, 6123, 6401, 6403, 6421, 6463, 6176, 6129, 6141, 6169, 6135, 6442, Series
1.9	IERC E2 Series (Extruded)
2.1	IERC E1, E3 Series (Extruded)
2.3 - 4.7	Wakefield 600 Series
4.2	IERC HP3 Series
4.5	Staver V3-5-2
5 - 6	IERC HP3 Series
5.2 - 6.2	Thermalloy 6103 Series
5.6	Staver V3-3-2
5.8 - 7.9	Thermalloy 6001 Series
5.9 - 10	Wakefield 680 Series
6	Wakefield 390 Series
6.4	Staver V3-7-224
6.5 - 7.5	IERC UP Series
8	Staver V1-5
8.1	Staver V3-5
8.8	Staver V3-7-96
8.8 - 14.4	Thermalloy 6013 Series
9.5	Staver V3-3
9.5 - 10.5	IERC LA Series
9.8 - 13.9	Wakefield 630 Series
10	Staver V1-3
13	Thermalloy 6117

Staver Co., Inc.: 41-51N. Saxon Ave., Bay Shore, NY 11706
IERC: 135 W. Magnolia Blvd., Burbank, CA 91502
Thermalloy: PO Box 34829, 2021 W. Valley View Ln., Dallas TX
Wakefield Engin Ind.: Wakefield, MA 01880

[1]All values are typical as given by mfgr. or as determined from characteristic curves supplied by mfgr.

α.) Next, determine B, which is approximately the height of the fins divided by 2, or 2.1 inches. Draw the second line from this value of B through $\alpha = 0.38$ and extend it all the way over to η. Now that we have found η graphically to be 86 percent, the equation for θ_{SA} can be solved.

If the solution for θ_{SA} is too large, this indicates an inadequate amount of heat-radiating ability; in such a situation, the fin height should be increased. If the equation yields a number that is less than the designed thermal resistance, 4 °C/W in this example, the heatsink is larger than it needs to be. A heatsink that is too large provides an additional thermal safety factor. But if the size of your case — or the cost of materials — prevents use of a large heatsink, you can choose a copper heatsink, a smaller one of the same material, or reduce the V_{IN} minus V_{OUT} differential. Any of these measures will help.

Returning to the derivation of θ_{SA}, h_r can also be improved upon from the 0.9 we used (for E) in the form of black anodized aluminum. **Table 2** indicates that black oil paint can have an E of 0.96. The preceding derivation of θ_{SA} assumed that the heatsink was vertically mounted and that the fins were symmetrically square. The maximum effectiveness of any heatsink can not be achieved with horizontal mounting; if a finned heatsink is mounted horizontally, this is even worse. If horizontal mounting is used, derate h_c by multiplying by 0.8. If the confines of the case necessitates horizontal mounting, the "one side only" finned heatsink should have its η multiplied by 0.5 and its h_c derated by being multiplied by 0.94.

Table 4. Use of silicon grease or adhesive decreases thermal resistance.

insulator	thermal resistance, °C/W dry	silicon grease
none	0.20	0.10
Teflon	1.45	0.80
mica	0.80	0.40
anodized aluminum	0.40	0.35

With the values used thus far, η has been graphically determined to be 0.86. Substituting our numbers in **eq. 5** we obtain the following:

$$\theta_{SA} = \frac{10^3}{2 \times 18 \times 0.86 \times 9.3} = 3.47°C/W$$

Incidentally, the 10^{-3} was changed to 10^3 by moving it to the numerator and changing the sign of the exponent. The value of 3.47°C/W is not exactly the desired 4.00°C/W we initially sought, but gives a margin of safety. Remember, when doing it yourself, these formulas apply to both round and square symmetrical fins. Rectangular 2:1 aspect ratio (length-to-height) fins may also be used with the appropriate deratings used. **Table 3** shows some typical θ_{SA} values per case size in commerically available heatsinks; by following **figs. 4** and **5** you can build your own, and know it will do the job!

guidelines for use

Avoid placing voltage regulators next to heat-generating components like power resistors. When using smaller package regulators like the TO-5, TO-39, or TO-92 cases, keep the lead lengths to a minimum and use the largest possible printed circuit board copper connector runs to provide a heat dissipation path for the regulator. Make sure that the heatsink-to-case surfaces mate. With larger heatsinks, this is more difficult, so for good thermal conduction, use a thin layer of silicon thermal grease such as Dow Corning 340, General Electric 662, or Thermalloy's "Thermacote" or a thermal adhesive. Metallic-oxide-filled silicon grease placed between two surfaces eliminates air gaps and fills in scratches on the two metal surfaces. (A typical transistor mounted dry with a 0.2°C/W resistance will exhibit half this resistance after a layer of thermal grease or adhesive is applied.)

Negative voltage regulators, in particular, require that the heatsink and metal voltage regulator case be electrically insulated from each other. To accomplish this, use either a fiberglass or mica 0.003 to 0.005-inch thick insulating washer. While the presence of the washer will naturally increase thermal resistance, this effect can be partially offset by applying thermal grease to each side of the washer. When mounting a regulator on a heatsink with fins, orient the fins in the vertical plane. And be *very* careful when bending the regulator's leads, should such action ever be necessary.

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ham radio

all-metal 2-meter J-pole antenna

5/8 wave vertical survives any weather

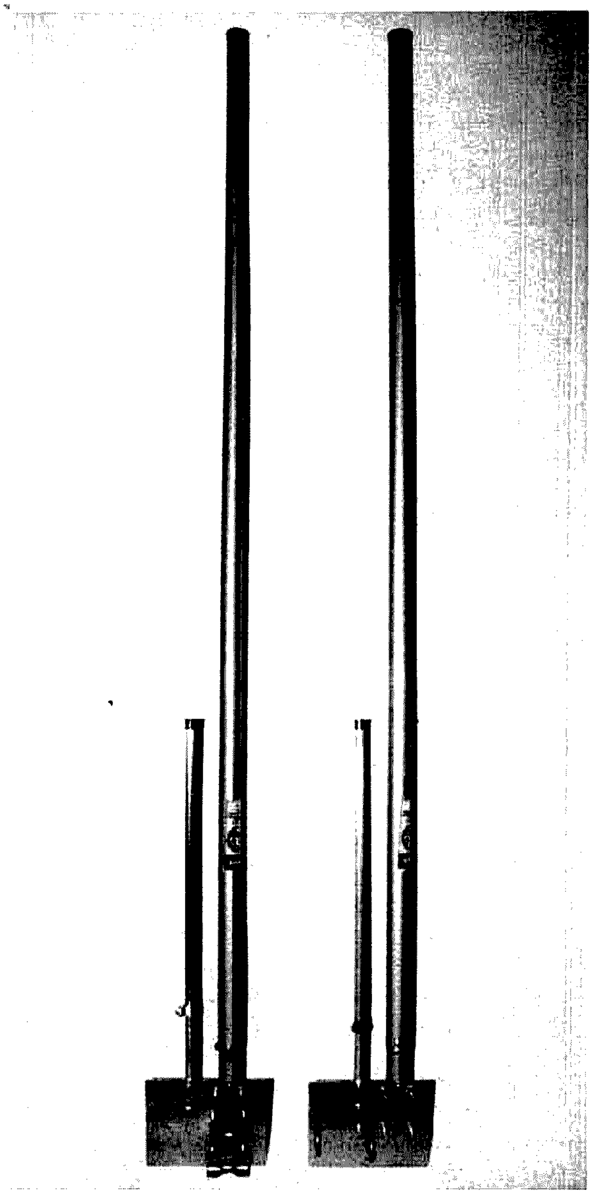

fig. 1. J-pole antenna: for upright mast mounting (left) and mounting to horizontal boom on tower (right).

While the J-pole antenna shown in **fig. 1** is by no means a new design,[1] it doesn't seem to be used often. Basically it's a 5/8-wave radiator with a 1/4-wave matching section. The matching section is positioned 1-1/4 inch from and parallel to the 5/8-wave element.

I've built two of these antennas; one is for mounting to a conventional upright mast, as shown by the unit on the left in **fig. 1**; the other is for mounting to a horizontal boom on a tower. No significant problems were experienced in either installation.

The advantages of the J-pole design are its omnidirectional radiation pattern, characteristic of vertical antennas, and its small-diameter installation area. It lacks the radials common to ground plane antennas, and does not require driven elements to be cut in order to tune for best match.

Advantages less readily apparent, but equally important, are the elimination of the need to insulate the radiating element from the ground system, making possible an all-metal unit of rigid construction. With the whole antenna effectively grounded and installed atop a tower or mast which is grounded at the base, there is very little static noise buildup. The additional gain possible over the quarter wave ground plane because of the increased length is considered yet another advantage.

construction

A list of materials is provided in **table 1**; dimensional information is shown in **fig. 2**. The mounting plate is shown in **fig. 3**. The extra set of four holes in the plate allows the J-pole to be mounted in either of two different planes, depending on the location of the final installation.

The total length of the longest piece of tubing is 57 inches (144.8 cm) from the tip to the mounting plate. This piece can be made from a 5-foot (152-cm) section of lightweight (20 gauge) mast available from Radio Shack. There is no reason to spend more for the heavier gauge mast because it doesn't support anything other than its own weight. (If you really want to spend more money on this antenna, put the money to good use and buy better coax, or stainless steel assembly hardware, or both.)

By Michael Hood, KD8JB, 849 Dickinson S.E., Grand Rapids, Michigan 49507

table 1. List of materials.

quantity	item
1	20-gauge galvanized steel TV mast. 5 ft. × 1-1/4 in. (152 × 3.18 cm)
1	length aluminum tubing, 22 × 1/2 × 0.035 inch (56 cm × 1.3 cm × 0.89 mm)
1	mounting plate, 6 × 4-1/2 × 1/8 inch (15.2 × 11.4 × 0.32 cm)
3	6-32 × 1 inch stainless steel panhead machine screws
3	No. 6 star lockwashers*
3	6-32 stainless steel hex nuts*
4	1-1/4 inch U-bolts
2	6-32 × 1/2-inch self-tapping sheet-metal screws**
1	1/2-inch diameter clamp (MS21919-DG5)
1	10-32 × 5/8-inch stainless steel panhead machine screw
2	No. 10 flat washers
2	No. 10 star lockwashers
2	10-32 stainless steel hex nuts
1	3/8 × 3 inch dowel
2	tubing caps for element ends (optional)

*Can be replaced by 6-32 nuts with star washers attached as part of the nut itself.

**One screw is not required if the through-bolt arrangement is used. In this case, a 10-32 × 1-1/2 inch cap screw, two star washers, two plain hexnuts, and one spring lockwasher will be required to replace the one 6-32 sheet-metal screw and its associated star washer. The choice is yours.

(Metric dimensions are approximate; build to fit. — Editor)

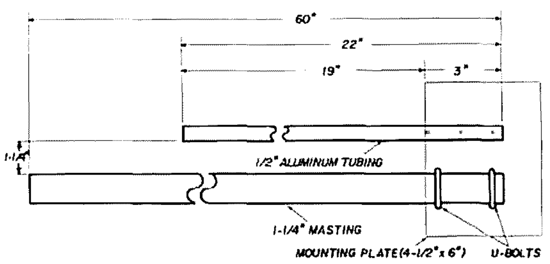

fig. 2. J-pole antenna dimensions.

The clamps holding the 5/8 wave section to the mounting plate are 1 to 1-1/4 inch (3.2 cm) U-bolts expanded slightly to pass comfortably around the compressed portion of the mast without distorting the tubing.

Electrical connection with the mounting plate is accomplished by filing or wirebrushing the paint from the bottom 3 inches (7.6 cm) of the mast where it comes in contact with the mounting plate. A second electrical connection is made by passing a self-tapping sheet-metal screw (6-32 or larger preferred) through the mounting plate and into the mast section as shown in **fig. 4**.

The quarter wave matching section is made from any gauge aluminum tubing of 1/2 inch (1.3 cm) O.D. In this case, 0.035 inch (0.89 mm) or so wall thickness was used; again, there's no need to buy heavier tubing because no weight is supported. A total of 19 inches (48.26 cm) must project above the mounting plate (see **fig. 2**), making the total length 22 inches (55.88 cm) for this portion of the assembly.

Electrical connection with the mounting plate is made by removing any oxidation from the tubing before bolting it to the mounting plate. Insert a piece of 3/8-inch (0.95 cm) wooden dowel into the end of the tubing until it's flush with the bottom edge; then bolt the matching section to the mounting plate with 6-32 × 1 inch machine screws, star lockwashers, and

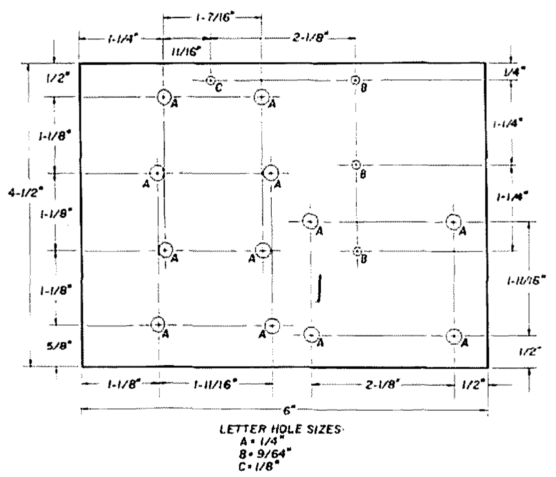

fig. 3. J-pole antenna mounting plate template.

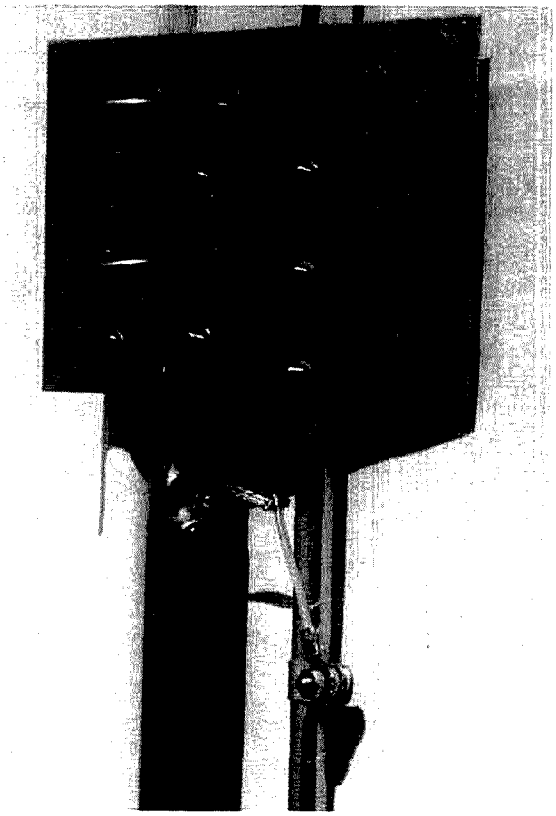

fig. 4. Assembly showing coaxial cable connections.

hexnuts (see **fig. 5**). Inserting the dowel permits the 6-32 hardware to be tightened without unduly distorting the tubing, which would allow hairline cracks to form and cause premature element failure.

The clamp used to attach the coax center conductor (**fig. 4**) is an aircraft-type Adel clamp with the rubber sleeve removed. This clamp was originally intended for 5/16-inch (0.79 cm) tubing, but with the rubber removed and the clamp installed as shown, a tight fit is maintained around the matching section. Any such device can be used as long as a good mechanical connection is maintained.

As shown in **fig. 4**, the shield of the cable is attached to the larger diameter element at a point 1-1/2 inches (3.8 cm) away from the mounting plate. Two different methods of attaching the shield to the 5/8-wave section were used; both are satisfactory, though I feel more confident about the No. 10 through-bolt in the 5/8ths section than I do about the No. 6 sheet-metal screw and lockwasher. This is a matter of personal preference. A No. 14 AWG terminal lug was used to terminate the shield as shown, but both lug ends were replaced with a solderable lug and then silver soldered on the completed installation.

If a certain amount of hardware "overkill" seems apparent in this project, it's because in the case of antennas, too much is usually better than too little, particularly where weather conditions are severe. It's a good idea to invest in stainless steel hardware and to protect the mounting plate assembly with a polyurethane coating or spray varnish, following the directions and precautions on the spray can. Do this *after* tuning the matching section.

tuning

To tune the matching section, start at a point about 2 inches (5 cm) from the mounting plate and apply RF at the frequency most likely to be used, for example, 146.52 MHz. Note the SWR at that frequency and adjust the clamp, sliding it as necessary to produce minimum SWR. The antennas in the photographs tuned at approximately 2-1/2 to 4 inches (6.4 to 10 cm) from the mounting plate. They did not tune to identical points on the respective matching sections. In both cases, the antennas tuned to an indicated SWR of 1.2:1 at 146.52 MHz. If you've worked with antennas before, you probably know that any SWR reading depends on so many factors that tuning for minimum SWR is, more often than not, sufficient for most purposes.

installation

Figure 6 shows the antenna mounted to a horizontal boom off the side of a mast. Convention dictates that the antenna point upward when installed, but there is no reason that it cannot be pointed downward. If high winds are a problem, or if the clamps used to attach the assembly to the boom are not of the gripper variety, pointing the antenna downward can help maintain its vertical position. As long as the antenna is mounted one or more wavelengths away from any large reflective or RF absorbent surface, fear of upside-down mounting is completely unfounded.

Conventional mounting to a vertical mast is accom-

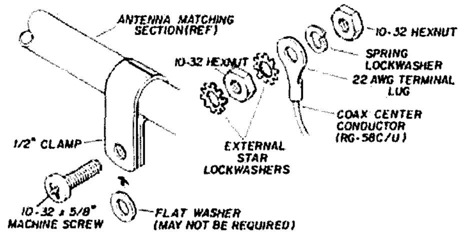

fig. 5. Half-inch clamp and hardware assembly sequence.

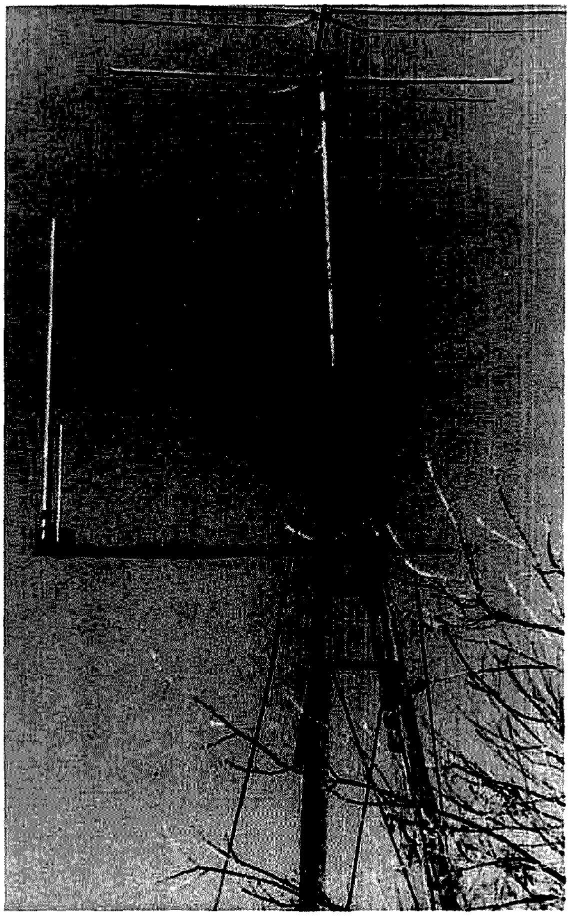

fig. 6. Horizontal mounting on a windmill tower.

plished by using the other set of mounting holes in the antenna mounting plate. No special problems were noted with this arrangement.

variations

As with any project, several additional ideas came to mind after the antennas were completed. For instance, the coaxial shield attachment to the 5/8th's element could certainly be made to be adjustable. Doing this might result in a better SWR reading, or perhaps allow the coax terminations to be physically matched more evenly at the antenna. If I were to build another, I would try this approach. Should you decide to make both connections adjustable, remember to remove the paint from the portion of the mast the clamp will rest against when tuned.

It may also be possible to make the 5/8th's element diameter smaller. The 1-1/4 inch tubing is convenient because it can be purchased at the proper length, but if you have stock left over from other antenna projects, it would be less expensive to use what you have. If you do, remember that antennas built from different stock may tune at positions other than those indicated in this article. I'd be delighted to hear from anyone who attempts this alternative approach.

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ham radio

wideband VCO design

Use this low-noise near-octave source in your HF receiver or VHF/UHF exciter

This article discusses the design and construction of a wideband voltage controlled oscillator (VCO). It can be used as an LO in a broadband receiver or spectrum analyzer, as a low-noise phase locked source in a VHF or UHF exciter, or alone as a basic signal generator.

This particular design is used in an HF communications receiver. When phase locked to a low-noise reference, it demonstrates excellent signal-to-noise performance — greater than 130 dBc/Hz at 20 kHz offset from the carrier. As mentioned in previous articles[1,2,3] this noise performance, or spectral purity of the oscillator, is quite important if the receiver is to perform well in the presence of strong adjacent channel signals.

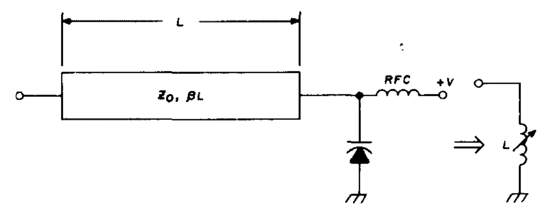

fig. 1. Transmission line and varactor are equivalent to tunable inductor.

design concept

The oscillator design uses a resonator or tank circuit which consists of a transmission line terminated by a voltage variable capacitor or varactor. A transmission line of correct length looks like a variable inductance as the varactor voltage is varied (**fig. 1**). The oscillator uses the Colpitts configuration, and takes advantage of the feedback capacitor voltage divider plus any parasitic capacitance present. If the Colpitts configuration is operated with a fairly high C to L ratio, then this condition tends to improve the oscillator operating Q and consequently the oscillator signal-to-noise ratio. To assist in the design of the VCO a BASIC computer program routine has been included. Some examples, as well as actual designs and performance data, are also provided. Finally some interesting applications are highlighted.

equations developed

The input impedance of a lossless transmission line, Z_s, is equal to:

$$Z_s = \frac{Z_\ell \cos \beta L + j Z_0 \sin \beta L}{\cos \beta L + j (Z_\ell / Z_0) \sin \beta L} \qquad (1)$$

If Z_ℓ is zero (the line is shorted at the end) from the preceding equation, $Z_s = j Z_0 \tan \beta L$ and the line has an inductive reactance of Z_0 ohms times the electrical length of the line. This is true provided the value of βL $(\beta = 2\pi/\lambda)$ is less than 90 degrees or the line is less than a quarter wavelength long. If the line is precisely a quarter wave long (electrically), then from eq. **1** we have:

$$Z_s = Z_0^2 / Z_\ell \qquad (2)$$

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In this case, if Z_ℓ were a capacitive reactance of $-jX_{C_v}$ then

$$Z_s = jZ_0{}^2/X_{C_v} \tag{3}$$

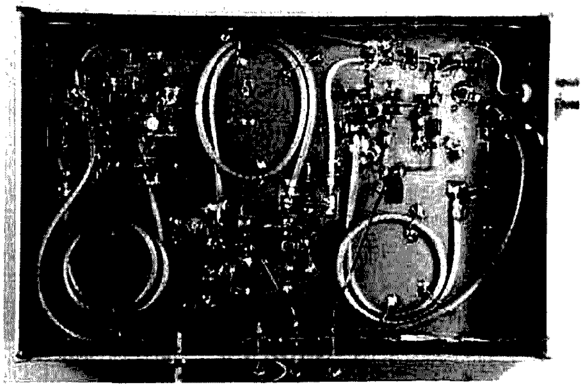
and the capacitive reactance of the termination is reflected to the input of the line as an inductive reactance of $j/X_{C_v} = Z_s/Z_0{}^2$ and an inductance of

$$L = C_v Z_o{}^2 \tag{4}$$

For example, an oscillator that tunes 135 MHz to 145 MHz requires the electrical length to remain a quarter wavelength or less at the maximum frequency of 145 MHz. If a 50-ohm transmission line is used, then the physical length of a quarter wave is calculated using:

$$\textit{length (feet)} \approx 246/f_0(MHz)\sqrt{\varepsilon_r} \tag{5}$$

where ε_r is the relative dielectric constant of the line. The ε_r for a 50-ohm line using a Teflon® dielectric is 2.2; therefore, the physical line length is 13.7 inches. If our oscillator includes device shunt capacitance, feedback capacitance, and parasitics totaling 20 pF, then the required inductance to achieve resonance is 60 nH at 145 MHz and 70 nH at 135 MHz. Consequently, the required variation in the varactor capacity is calculated from eq. 4 as 24 pF to 28 pF or 4 pF. If a lower Z_0 line were chosen, and the total capacitance remained the same, an increase in the varactor capacitance change would be required.



Switched transmission line injection oscillators as shown in fig. 2.

The equations developed so far allow us to choose a value for the transmission line impedance Z_0 and obtain a value of inductance L for a given varactor capacity, C_v. This equivalent inductance then becomes part of our oscillator tank. Since a transmission line and capacitor are readily controlled, it becomes very attractive to construct VHF and UHF oscillators in this manner since small fixed inductances are difficult to control. Furthermore, the resonator or tank is a shielded structure and therefore quite immune to outside disturbances, such as, microphonics, which can lead to hum and modulation effects.

three parameters affect frequency

Now consider what happens if the transmission line length, impedance, or the oscillator shunt capacitance is varied. Under these conditions, any value of inductance necessary to resonate the oscillator tank can be developed. Given these three variables, an oscillator frequency change of 50 percent can be achieved for a given value of transmission line length and impedance Z_0. With careful design, even an octave is possible. The necessary design equations are presented in the **appendix**, and a BASIC program is included to solve for the transmission line impedance, Z_0, electrical length, and required varactor capacitance change. Finally, the transmission line *physical* length is calculated, with the dimensions given in both inches and metric units. Because the program is in-

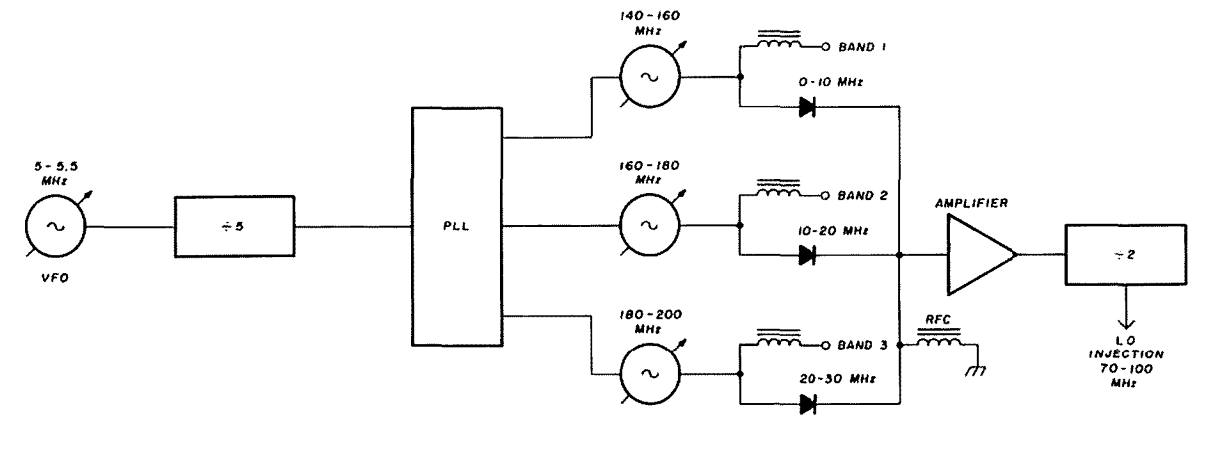


fig. 2. HF communications receiver using switched transmission line VCOs.

Program 1. Varactor oscillator design program listing.

```
]LIST

10  HOME : REM CLEARS THE SCREEN
20  PRINT "  VARACTOR TUNED WIDEBAND"
30  PRINT "   OSCILLATOR DESIGN"
40  PRINT "        A.M.VICTOR"
50  PRINT "        1.1 9/9/84"
60  PRINT : PRINT : PRINT
70 PI = 3.1415926
80  PRINT "WHAT IS THE MIN AND MAX TUNE": PRINT "RANGE IN MHZ ";: INPUT
     F1,F2
90 W2 = 2 * PI * F2 * 1E6
100 W1 = 2 * PI * F1 * 1E6
110  PRINT : PRINT "ENTER TRANSMISSION LINE ZO,": PRINT "AND STARTING LI
     NE LENGTH": PRINT "IN DEGRESS";: INPUT ZO,ITHETA
120 STHETA = ITHETA
130  PRINT : PRINT "ENTER LINE LENGTH INCREMENT";: INPUT DL
140  PRINT : PRINT "ENTER THE DEVICE CAPACITANCE": PRINT "PLUS ANY PARAS
     ITICS IN PF";: INPUT CIN
150  HOME : REM BEGIN CALCULATIONS
160 XIN = 1 / (W1 * CIN * 1E - 12)
170 X2IN = 1 / (W2 * CIN * 1E - 12)
180  PRINT : PRINT "DELTA XCV "; TAB( 12);"LENGTH"; TAB( 20);"CVMIN"; TAB(
     30);"CVMAX"
190  FOR THETA = ITHETA TO 180 STEP DL
200 RTHETA = THETA / 57.29578
210 A = (ZO *  TAN ((W2 / W1) * RTHETA)) - (W1 / W2) * XIN
220 B = 1 + (XIN *  TAN (RTHETA) / ZO)
230 C = 1 + ((W1 / W2) * (XIN / ZO) *  TAN ((W2 / W1) * RTHETA))
240 D = ZO *  TAN (RTHETA) - XIN
250 XQUOT = (A * B) / (C * D)
260 XCVW1 = (ZO *  TAN (RTHETA) - XIN) / (1 + (XIN / ZO) *  TAN (RTHETA)
     )
270 Y2CVW2 = ((ZO *  TAN ((W2 / W1) * RTHETA) - X2IN)) / (1 + (X2IN / ZO
     ) *  TAN ((W2 / W1) * RTHETA))
280 CMAX = 1 / (2 * PI * F1 * XCVW1 * 1E6)
290 CMAX = CMAX * 1E12
300 X = CMAX
310  GOSUB 450
320 CMAX = H
330 YCMIN = 1 / (2 * PI * F2 * Y2CVW2 * 1E6)
340 YCMIN = YCMIN * 1E12
350 X = YCMIN
360  GOSUB 450
370 YCMIN = H
375  REM : NEXT 4 LINES CHECK FOR VALID TRANSMISSION LINE LENGTHS AND CAP
     ACITOR VALUES.
380  IF XQUOT < 0 THEN 430
390  IF YCMIN < 0 THEN 430
400  IF CMAX < 0 THEN 430
410  IF YCMIN > CMAX THEN 430
420  PRINT XQUOT; TAB( 13);THETA;"DEG"; TAB( 20);YCMIN;"PF"; TAB( 30);CM
     AX;"PF"
430  NEXT THETA
440  GOTO 530
450 G =  INT (X)
460 H = G - X
470 H =  ABS (H)
480 H = 100 * H
```

```
490 H =  INT (H)
500 H = H / 100
510 H = H + G
520  RETURN
530  PRINT "ANOTHER RUN WITH NEW PARAMETERS Y/N ";: INPUT R$
540  IF R$ = "N" THEN 740
550  HOME : PRINT : PRINT "EDIT WHICH PARAMETERS?"
560  PRINT "FREQUENCY RANGE (1)?"
570  PRINT "TRANSMISSION LINE (2)?"
580  PRINT "DEVICE CAPACITANCE (3)?"
590  PRINT "OR DO TRANSMISSION LINE LENGTH CALCULATION (4)?"
600  PRINT "START NEW CALCULATION (5)"
610  PRINT "OR (6) TO END"
620  INPUT T
630  ON T GOTO 640,680,700,750,150,850
640  PRINT "ENTER TUNING RANGE MIN,MAX";: INPUT F1,F2
650 W1 = 2 * PI * F1 * 1E6
660 W2 = 2 * PI * F2 * 1E6
670  GOTO 550
680  PRINT "ENTER TRANSMISSION LINE ZO";: INPUT ZO
690  GOTO 550
700  PRINT "ENTER DEVICE CAPACITANCE PLUS ANY PARASITICS IN PF";: INPUT
     CIN
710  GOTO 550
720  HOME :ITHETA = STHETA
730  GOTO 160
740  HOME
750  PRINT "TRANSMISSION LINE LENGTH CALCULATION"
760  PRINT : PRINT "ENTER REL.DIELECTRIC CONSTANT OR": PRINT "VELOCITY F
     ACTOR, (E) OR (V) AND VALUE";: INPUT K$,ER
770  IF K$ = "V" THEN ER = 1 / ER ^ 2
780  PRINT "ENTER TRANSMISSION LINE": PRINT "LENGTH IN DEGREES";: INPUT
     FTHETA
790 LM = (FTHETA / 360) * (300 / F1) * (1 /  SQR (ER))
800 IM = (LM * 100) / 2.54
810  PRINT "METRIC LENGTH"; TAB( 30);"INCHES"
820  PRINT LM; TAB( 30);IM
830  PRINT "ANOTHER RUN Y/N";: INPUT R$
840  IF R$ = "Y" THEN 550
850  END
```

teractive, you can try various combinations of line impedance, line length, and oscillator shunt output capacitance, which can be altered by shunting the oscillator device with an external capacitance. The program is arranged to automatically increment the transmission line length by a specified amount. During the program operation you'll notice that a particular line length exists which minimizes the required varactor capacitance change. Depending on the oscillator's tuning range and the line impedance, the required capacitance change of the varactor is affected and changes. For a large percentage tuning bandwidth, a high Z_0 is needed to keep the varactor capacitance change within practical limits.

VCO circuit

Several versions of the oscillator were tried. The VCOs shown in **fig. 2** and in the photograph work well in an HF receiver using up-conversion. Each of the three VCOs covers 20 MHz in the 140-200 MHz range. The oscillators are phaselocked to a 1 MHz reference and their outputs are divided by 2. The final output is then tunable from 70 to 100 MHz in 500 kHz steps and provides injection for an up-conversion receiver with a 70 MHz first IF. U310 J-FETs and MV109 varactors are used. Trimmer capacitor values, varactor capacitance swing, and transmission line dimensions were determined by using the computer program. Each oscillator uses clamp bias to help regulate oscillator feedback as it is tuned and to prevent the FET gate-source junction from becoming forward-biased. This appears to help the oscillator noise performance as evidenced by measured values of better than 135 dBc/Hz, measured 25 kHz away from the car-

Program 2. Varactor oscillator design examples.

```
ANOTHER RUN WITH NEW PARAMETERS Y/N ?N
TRANSMISSION LINE LENGTH CALCULATION

ENTER REL.DIELECTRIC CONSTANT OR
VELOCITY FACTOR, (E) OR (V) AND VALUE?E,2.2
ENTER TRANSMISSION LINE
LENGTH IN DEGREES?45
METRIC LENGTH                 INCHES
.252824948                    9.95373812
ANOTHER RUN Y/N?Y

EDIT WHICH PARAMETERS?
FREQUENCY RANGE (1)?
TRANSMISSION LINE (2)?
DEVICE CAPACITANCE (3)?
OR DO TRANSMISSION LINE LENGTH CALCULATION (4)?
START NEW CALCULATION (5)
OR (6) TO END
?1
ENTER TUNING RANGE MIN,MAX?200,400

EDIT WHICH PARAMETERS?
FREQUENCY RANGE (1)?
TRANSMISSION LINE (2)?
DEVICE CAPACITANCE (3)?
OR DO TRANSMISSION LINE LENGTH CALCULATION (4)?
START NEW CALCULATION (5)
OR (6) TO END
?5

DELTA XCV  LENGTH  CVMIN    CVMAX
164.421688  15DEG  12.84PF  4223.22PF
7.06193955  20DEG  8.34PF   117.86PF
5.11163621  25DEG  5.8PF    59.3PF
4.8057213   30DEG  4.07PF   39.2PF
5.22462976  35DEG  2.76PF   28.92PF
6.73458529  40DEG  1.67PF   22.6PF
12.9839108  45DEG  .7PF     18.27PF
ANOTHER RUN WITH NEW PARAMETERS Y/N ?N
TRANSMISSION LINE LENGTH CALCULATION

ENTER REL.DIELECTRIC CONSTANT OR
VELOCITY FACTOR, (E) OR (V) AND VALUE?V,.66
ENTER TRANSMISSION LINE
LENGTH IN DEGREES?40
METRIC LENGTH                 INCHES
.11                           4.33070866
ANOTHER RUN Y/N?N

]RUN
  VARACTOR TUNED WIDEBAND
   OSCILLATOR DESIGN
        A.M.VICTOR
        1.1 9/9/84
```

```
WHAT IS THE MIN AND MAX TUNE
RANGE IN MHZ ?100,200

ENTER TRANSMISSION LINE ZO,
AND STARTING LINE LENGTH
IN DEGRESS?50,10

ENTER LINE LENGTH INCREMENT?5

ENTER THE DEVICE CAPACITANCE
PLUS ANY PARASITICS IN PF?40

DELTA XCV   LENGTH   CVMIN     CVMAX
62.3339054  40DEG    9.82PF    1225.29PF
22.0994892  45DEG    6.33PF    279.89PF
23.7659259  50DEG    3.29PF    156.62PF
114.03646   55DEG    .47PF     107.54PF
ANOTHER RUN WITH NEW PARAMETERS Y/N ?Y

EDIT WHICH PARAMETERS?
FREQUENCY RANGE (1)?
TRANSMISSION LINE (2)?
DEVICE CAPACITANCE (3)?
OR DO TRANSMISSION LINE LENGTH CALCULATION (4)?
START NEW CALCULATION (5)
OR (6) TO END
?2
ENTER TRANSMISSION LINE ZO?75

EDIT WHICH PARAMETERS?
FREQUENCY RANGE (1)?
TRANSMISSION LINE (2)?
DEVICE CAPACITANCE (3)?
OR DO TRANSMISSION LINE LENGTH CALCULATION (4)?
START NEW CALCULATION (5)
OR (6) TO END
?5

DELTA XCV   LENGTH   CVMIN     CVMAX

28.0326343  30DEG    10.55PF   591.88PF
11.6044757  35DEG    7.38PF    171.5PF
10.1093855  40DEG    4.91PF    99.38PF
12.2899117  45DEG    2.81PF    69.17PF
29.0553268  50DEG    .9PF      52.38PF
```

rier frequency (see **fig. 3**). Each VCO output is combined through switched pin diodes and the VCO is enabled by switching DC power from a bandswitch control logic board. The tuning line voltage of each VCO is tied together through shielded coax and 400 nH RF chokes. A small series resistor is used in line with the chokes to provide for a lower choke *Q*. Note that a small resistor is used; large decoupling resistors add excessive amounts of thermal noise and can ruin an otherwise low-noise design.

digital VFO applications

The octave-tuned oscillator presented in **fig. 4** covers 100-200 MHz. The layout is more critical since any additional parasitic capacitance present tends to reduce the achievable tuning range. The circuit does

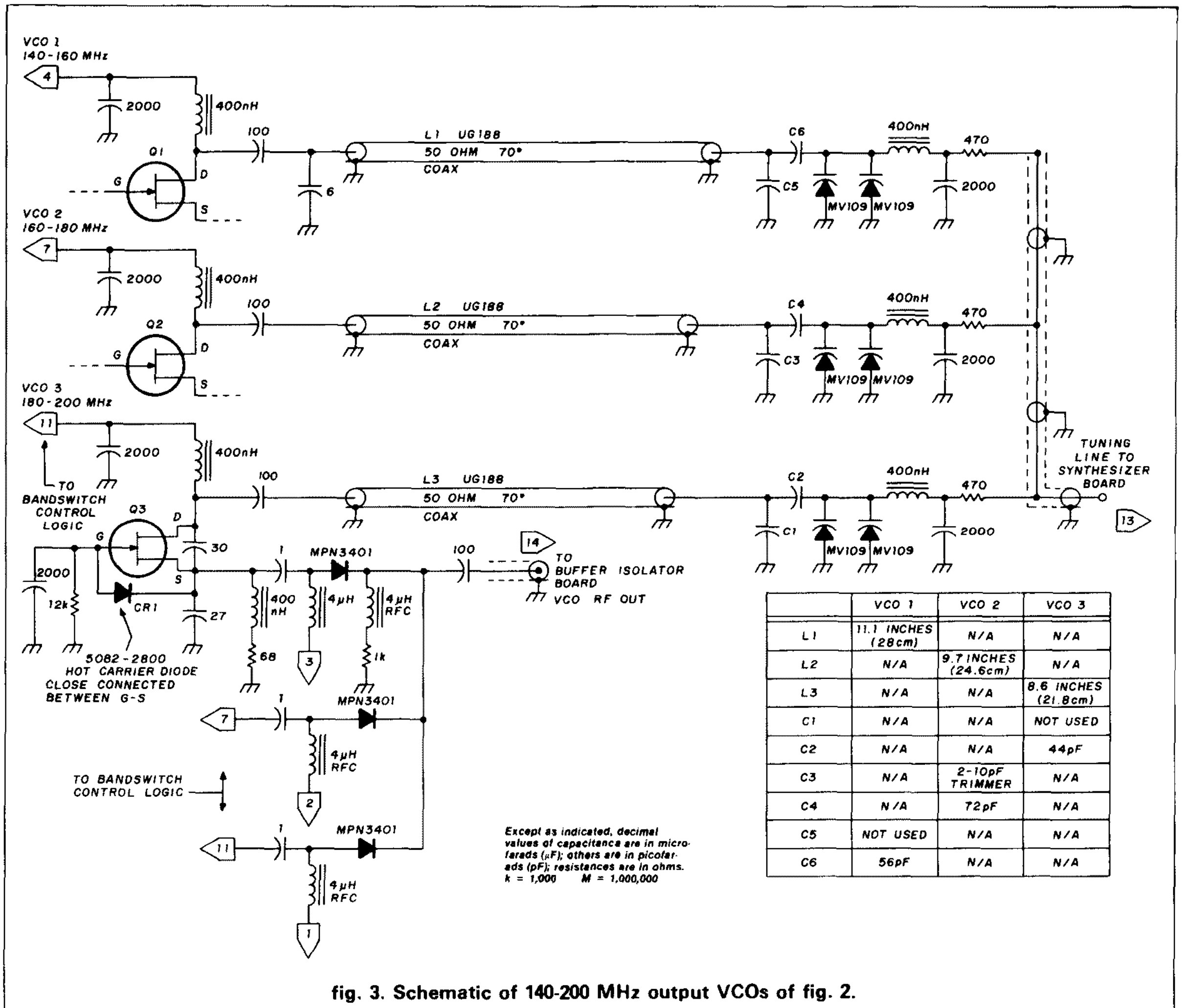
provide octave coverage and requires a varactor voltage change from 1 to 20 volts. The noise performance was less impressive (115 to 120 dBc/Hz). However, if used in a low-noise phase lock loop and followed by frequency division, it provides the basis for a low noise synthesizer or digital VFO with 10 Hz resolution.

Another example of a digital VFO is shown in **fig. 5**. How about a continuous-tuning HF receiver with 1 kHz steps, using a single phase lock loop and the popular 9 MHz IF? The approach outlined uses a single 200 to 400 MHz VCO. Despite the coverage of the VCO, which is only an octave, appropriately dividing down the VCO frequency provides for frequency coverage in *excess* of an octave! If the first IF is 9 MHz, then an LO injection frequency from 10 to 40 MHz will provide reception from 1 to 31 MHz. Using the 200 to 400 MHz VCO with a divide by 10/20 circuit following it yields the needed coverage. The VCO is locked to a 10 kHz reference and programmed in 1 kHz steps. Steps smaller than 1 kHz could be provided for by tuning or synthesizing the BFO.

A 200 to 400 MHz VCO was constructed and the measured noise performance varied from 110 dBc/Hz to 115 dBc/Hz at 25 kHz offset from the carrier. Division by 10 or 20 improves the noise by at least 20 dB or 130 dBc/Hz, which is quite adequate for most receiver applications. Signal generation is handled in a similar fashion, using a cascade of binary dividers. Our 200 to 400 MHz VCO, could provide the basis of an HF through UHF source by using successive division of the VCO and bandpass filtering. A number of commercially available signal sources use this approach.

summary

A design technique which allows the construction of wideband tunable oscillators with excellent noise performance is presented. The technique uses a trans-



	VCO 1	VCO 2	VCO 3
L1	11.1 INCHES (28cm)	N/A	N/A
L2	N/A	9.7 INCHES (24.6cm)	N/A
L3	N/A	N/A	8.6 INCHES (21.8cm)
C1	N/A	N/A	NOT USED
C2	N/A	N/A	44pF
C3	N/A	2-10pF TRIMMER	N/A
C4	N/A	72pF	N/A
C5	NOT USED	N/A	N/A
C6	56pF	N/A	N/A

fig. 3. Schematic of 140-200 MHz output VCOs of fig. 2.

mission line and voltage variable capacitor that simulates an inductive change. A BASIC program which aids in the design of this configuration is included. Oscillators which exhibit broadband and low-noise performance have many applications. I would be very interested in hearing from readers who have similar applications.

references

1. Doug DeMaw and Wes Hayward, "Modern Receivers and Transceivers: What Ails them?" *QST*, January, 1982, pages 13-14.
2. William F. Egan, *Frequency Synthesis By Phaselock*, John Wiley and Sons, 1981, pages 196-201.
3. D.B. Leeson, "A Simple Model of Feedback Oscillator Noise Spectrum," *Proceedings of the IEEE*, February, 1966, page 329.
4. Ulrich L. Rohde, *Digital PLL Frequency Synthesizers*, Prentice-Hall, Inc., 1983.
5. W.H. Hayward, *Introduction to Radio Frequency Design*, Prentice-Hall, Inc., 1982.

appendix

determination of the transmission-line resonator parameters

The transmission line formula (**eq. 1**) can be rearranged to solve for Z_0. Note that βL is not restricted to an electrical length of 90 degrees or a quarter wavelength.

$$Z_\ell = \frac{JZ_0 \tan \beta L - Z_s}{j(Z_s/Z_0) \tan \beta L - 1} \quad \textbf{(A1)}$$

Now $Z_s = +jX_{in}$ and resonates with $-jX_{in}$ of the device at resonance. Therefore, substituting for Z_s in **eq. A1** gives:

$$Z = \frac{-j(Z_0 \tan \beta L - X_{in})}{1 + \dfrac{X_{in} \tan \beta L}{Z_0}} \quad \textbf{(A2)}$$

and the reactance of our tuning varactor is:

$$X_{C_v} = \frac{Z_0 \tan \beta L - X_{in}}{1 + \dfrac{X_{in} \tan \beta L}{Z_0}} \quad \textbf{(A3)}$$

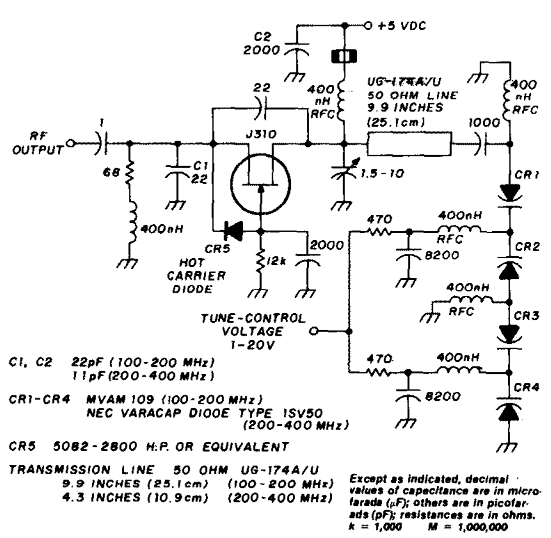


fig. 4. A pair of octave tuned oscillators 100-200 MHz or 200-400 MHz.

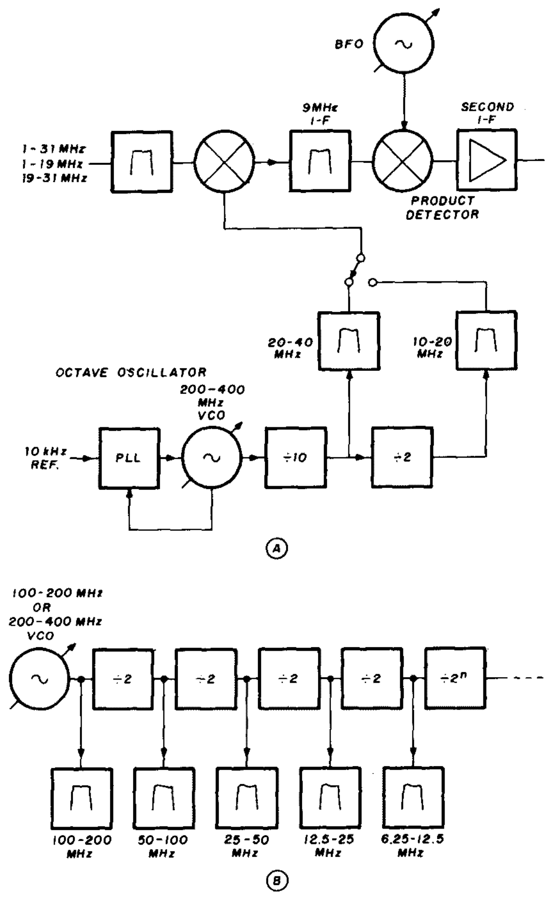


fig. 5. (A) a continuous tuning HF receiver or (B) a general purpose signal source using binary stepped band selection.

Simplify by setting $\beta L = X$ at frequency f_1, and $\beta L = \frac{f_2}{f_1}X$ at frequency f_2. Since the oscillator shunt capacitance remains constant, X_{in} at f_2 equals (f_1/f_2) times X_{in} at f_1. This information is now used to obtain the following reactance change needed to tune the VCO from f_1 to f_2. The following expression can be solved using the BASIC computer program or even a hand-held calculator. The equation is:

$$\frac{X_{C_v}(f_2)}{X_{C_v}(f_1)} = \frac{\left[Z_0 \tan\left(\frac{f_2}{f_1}X\right) - \frac{f_1}{f_2}X_{in}(f_1)\right]\left[1 + \frac{X_{in}(f_1)\tan X}{Z_0}\right]}{\left[1 + \frac{f_1}{f_2}\cdot\frac{X_{in}(f_1)}{Z_0}\tan\left(\frac{f_2}{f_1}X\right)\right]\left[Z_0 \tan X - X_{in}(f_1)\right]}$$

where $X_{in}(f_1)$ = oscillator total shunt reactance at frequency f_1 **(A4)**

$X_{in}(f_2)$ = oscillator total shunt reactance at frequency f_2

$X_{C_v}(f_1)$ = varactor reactance at frequency f_1

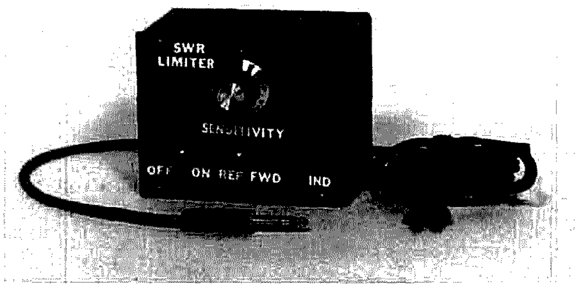
$X_{C_v}(f_2)$ = varactor reactance at frequency f_2

Using the BASIC program and **eq. A4** values of X, Z_0, X_{in} will yield the required tuning ratio.

ham radio

the weekender

reflected power limiter



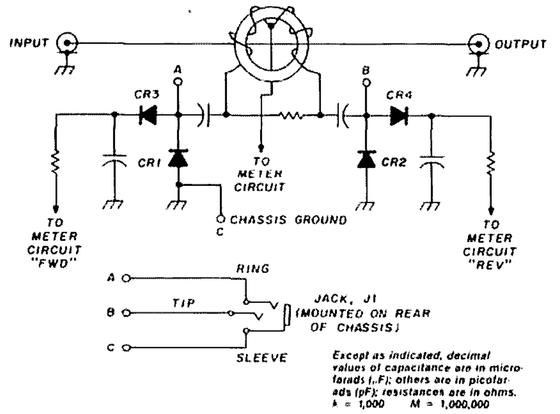


fig. 1. Partial SWR meter circuitry to show jack/diode connections.

Most of the newer transistorized transceivers have circuits that limit output power if a high SWR or reflected power condition is encountered. But because many earlier rigs — and also linears — do not have such protection, the following unit was designed and built to automatically shut down equipment if a pre-set value of reflected power is exceeded. As some of us already know through sad experience, high reflected power, even if only for a few seconds, can be both damaging and costly.

operation

The idea is to sample a small amount of RF voltage from your SWR or reflected power meter, rectify it, and feed it through a sensitivity control to the base input of a Darlington transistor. At the proper triggering level, a relay in the collector circuit will close two of its contacts, locking itself in. At the same time another set of relay contacts will open the AC input power to your transceiver, shutting it off. The RF voltage selected by switch S2 is either REFLECTED or FORWARD, which makes it easy to initially set up the unit. A switch, S3, provides an override circuit which simplifies initial setup.

A Darlington transistor was chosen because it has a Beta (current gain) of more than 5000, which provides more than sufficient sensitivity. The LED is useful in setting the bias control, and also indicates when the relay is locked in. During the breadboard development of the unit, some troublesome RF and 60 cycle pickup was encountered. This was readily cured by adding capacitors C1 and C2 as shown. They were incorporated into the final finished unit, as a precaution to eliminate such problems.

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My own RF wattmeter is a SWAN WM-1500, and is generally representative of how SWR meters and wattmeters are constructed. The partial schematic is shown to indicate how the FORWARD and REFLECTED voltages are picked off and rectified by the added diodes CR1 and CR2. A stereo jack, J1, grounded to the meter chassis is a neat, easy way to transfer the voltages to the unit. **Fig. 1** shows the schematic of the pick off voltages. At first glance it might seem that the DC voltage required for the unit could be picked off directly at the outputs of CR3 and CR4, thus eliminating the need for the added diodes CR1 and CR2. This worked well, except that under certain circumstances, a small amount of DC voltage from the unit would feed back into the wattmeter and cause the indicating meter of the wattmeter to read downscale by a few microamperes. The addition of the two diodes as shown eliminated that problem.

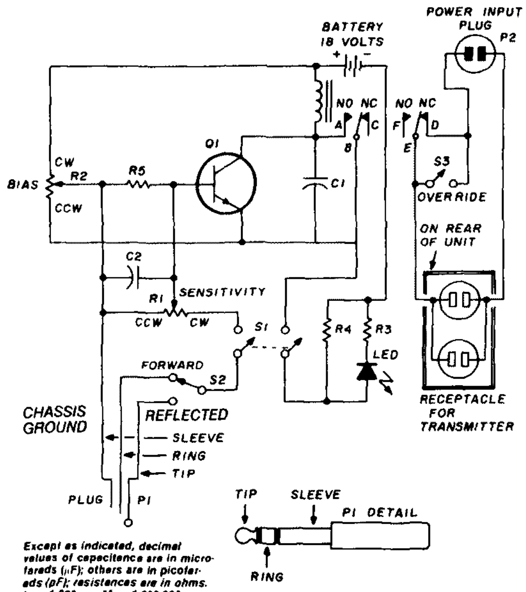
Without any RF voltage being fed into the unit, the bias control, if turned up sufficiently in the clockwise position, will turn on the transistor, locking in the relay. The bias control is a screwdriver adjustment on the rear of the unit. (It could also have been more conveniently mounted on the panel, using a conventional knob.) It is normally set somewhat below the turn-on point of the LED and relay lock in. The lower the bias control is set, the lower the standby current will be; I found that for normal operation, the standby current is less than a half a milliampere, which makes for long battery life. Switch S1 performs two functions, turning the power to the unit on, and when in the off position, isolating the unit from the wattmeter. Actually, the loading effect is so small as to be almost undetectable; the circuit was added more as a precaution than as a necessity.

The easiest way to set up the unit is to first determine the level of reflected power at which you want your transmitter to turn OFF. With the override switch, S3, on, and the bias properly set as just described, turn your sensitivity control counterclockwise. Set switch S2 in the FORWARD position. Then load up your transmitter until the forward power is the amount at which you want your reflected power level to trip the relay. Now turn the sensitivity control (clockwise) until the LED lights and the relay locks in. Throw S2 to the REFLECTED position, reset the relay by momentarily turning switch S1 OFF and then ON, and turn the override switch S3 OFF. The unit is now calibrated and ready to go. Now, if your reflected power in your system reaches the preselected level already set up, your LED will light up, your relay will lock in, and contacts DE of the relay will open, turning off your transmitter almost instantaneously and automatically protecting your equipment.

The required level of reflected power can be easily determined by knowing that if it is one tenth of the maximum forward power, then your SWR will be slightly below 2:1. Naturally, each operator may want to determine the maximum SWR at which the equipment will be shut off. When the unit is set up for its most sensitive condition, it actually will trip at about 3 to 4 watts of reflected power, which I have found is more sensitivity than you will normally need. Although the unit and the circuit shown in **fig. 2** turn off my transceiver input power, the relay contacts could instead easily be used to reduce your drive, increase your bias, reduce your plate voltage, or any one of the many different ways to reduce power output. The unit shown was built for my Yaesu FT-101-B transceiver. Its relay contacts have a current capacity of 3 amperes. Some experimentation was also done using an NPN Darlington transistor TIP 120, (Radio Shack™ part number 276-2068), which has a power dissipation of 65 watts, and would therefore be suitable for a high power linear control, in which a larger relay with heavier current contacts would be needed. It works in the circuit shown, but is slightly less sensitive. None of the parts listed were critical in any way; most of them came from my collection gathered over more than fifty years of hamming. The unit is housed in a metal box measuring 4 × 5 × 3 inches (10 × 12.7 × 7.6 cm). Any similar metal box will do.

I'd be interested in hearing from others who have built or experimented with a similar unit, and will be glad to answer any questions I can. Just include an SASE with your letter, addressed to me at the address provided.

ham radio



Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

item	description
C1	1 μF capacitor
C2	0.01 μF capacitor
CR1-CR4	IN34 diodes
J1	stereo jack
LED	miniature red LED Radio Shack™ No. 276-026
P1	stereo plug
P2	115 VAC plug
Q1	Darlington transistor, MPS-A13 Radio Shack No. 276-2060
R1	100 kilohm 2-watt linear potentiometer, carbon
R2	50 ohm 2-watt linear potentiometer, carbon
R3	390 ohm 1/2-watt carbon resistor
R4	100 kilohm 1/2-watt carbon resistor
R5	100 kilohm 1/2-watt carbon resistor
relay	12 VDC DPDT, Radio Shack No. 275-206A
rec.	two-outlet power receptacle
S1	DPST toggle switch, miniature size
S2	SPDT toggle switch, miniature size
S3	SPST toggle switch 250 volt, 6 ampere
batteries	two 9-volt, in series, Radio Shack 23-464

fig. 2. Reflected power limiter schematic.

applied Yagi antenna design part 3: 432 MHz with Knadle and Tilton

Computer model analyzes, optimizes Knadle, Tilton/Greenblum antenna designs

From the viewpoint of the antenna experimenter, the 432-MHz band is one of Amateur Radio's more intriguing frequency assignments. A Yagi with a boomlength of many wavelengths can be easily constructed, and stacking distances allow almost any operator to build arrays with formidable levels of performance. But with these advantages come several possible pitfalls: feedline losses can be a considerable factor in how well an array performs, and even the best of coaxial cables are noticeably lossy at 432 MHz. The matching system is very critical, and great care must be taken to avoid power losses between the driven element and the feedline.

While greater precision is required in cutting the parasitic elements for a 432 MHz Yagi than for a 144 MHz Yagi, precise element lengths are not as critical as is commonly believed. As is obvious from even a cursory examination of the tables of antenna iterations presented in this article, considerable latitude in the precise lengths of a series of parasitic elements is possible before the operator at either end of the signal path would notice any difference in signal strength. Finally, even at today's inflated prices, a single pound of welding rod would yield the elements for many long Yagis at 432 MHz.

432 MHz computer iterations

Two different Yagi design approaches — based on the Knadle and Tilton/Greenblum design data — are discussed in this article. Both are familiar to VHF/UHF Amateurs; both enjoy good reputations for excellent performance and reproducibility; and both resulted from long hours of effort on the parts of well known, experienced Amateurs. The design frequency is 432.0 MHz. Previous articles used 144.5 MHz and 220.5 MHz for those respective VHF bands.[1,2] This was a function of the stated design frequencies used in the development of the Yagis being iterated, or of the specific usage patterns of a particular band's frequency assignments. As is true of these two bands, the majority of weak-signal activity on 432 MHz is very close to 432.0 MHz. In addition, even long Yagis are comparatively broadbanded in frequency response on 432 MHz, making the use of 432.1 or 432.5 MHz as a design frequency of little practical purpose. An 8 MHz bandwidth, 428-436 MHz, is used to provide frequency response parameters. Because 0.0625 inch is probably the smallest measurement to which most Amateurs can accurately cut, it is used as the iteration increment.

the K2RIW Yagi antenna

Richard Knadle, K2RIW, is one of the most prolific contributors to Amateur VHF/UHF practice; his anten-

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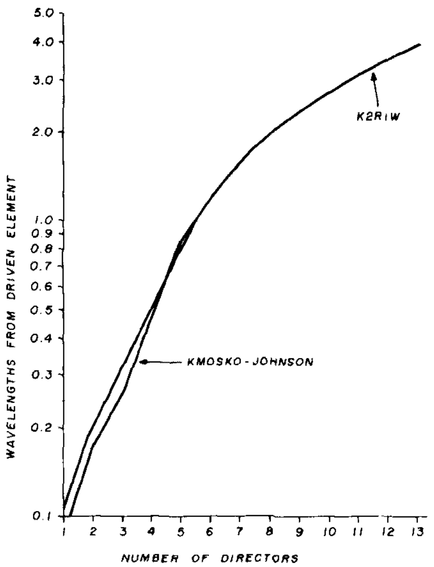

fig. 1. Comparison between the Kmosko-Johnson and Knadle 432 MHz Yagis — boomlength (starting from driven element) versus number of directors.

na and EME work is in addition to the transmitters, amplifiers, and receiver pre-amplifiers whose designs he has published in Amateur Radio journals. His original Yagi antenna design has been modified for introduction as a commercial product.[3] Yet another modification of his original design is the subject of this article's computer iterations.

For a while it seemed that locating a copy of the Knadle-Yagi antenna design was akin to trying to locate a schematic of the fabled Black Widow VHF/UHF transceiver. However, a full set of specifications was finally found in a rather obvious place,[4] and the optimization process was begun.

The possible origin of the K2RIW Yagi may be apparent from **table 1** and **fig. 1**. While its similarity to the Kmosko-Johnson design is apparent, the basic Knadle antenna has slightly more forward gain than the basic Kmosko-Johnson antenna (15.657 dBi versus 15.601 dBi). These basic antennas have thirteen elements and a zero director taper. What is perhaps the most interesting result of baselining the original K2RIW Yagi was finding the peak performance to be calculated to occur at the exact parasitic element lengths Knadle specified. Such close agreement between the empirical findings of a prominent practitioner and the iterations of a mathematical model is always a source of joy to the model's developers and users.

Before presenting the results of the computer iterations, it is worth noting that Knadle's exact matching method should be used on K2RIW Yagis. Other methods are probably equally efficient,[5] but by following the designer's instruction for at least the first antenna, the builder is able to establish a baseline for later comparisons.

table 1. A comparison of the interelement spacings of the Knadle and Kmosko-Johnson thirteen-element Yagis.

element		spacing in wavelengths from previous element	
number	name	Knadle	Kmosko-Johnson
1	reflector	0.0000	0.0000
2	driven	0.2289	0.2449
3	director 1	0.1098	0.0857
4	director 2	0.0984	0.0918
5	director 3	0.1235	0.0918
6	director 4	0.2013	0.1959
7	director 5	0.2745	0.3918
8	director 6	0.4323	0.3918
9	director 7	0.3889	0.3918
10	director 8	0.3889	0.3918
11	director 9	0.3889	0.3918
12	director 10	0.3889	0.3918
13	director 11	0.3889	0.3918

table 2. Baselined K2RIW Yagi at 432.0 MHz with fixed parasitic element spacings and parasitic element lengths supplied during each iteration.

element		length	element spacing	cumulative length
number	name	(inches)	(λ)	(λ)
1	reflector	—	0.000000	0.000000
2	driven	12.991445	0.228755	0.228755
3	director 1	—	0.109802	0.338557
4	director 2	—	0.098364	0.436922
5	director 3	—	0.123527	0.560450
6	director 4	—	0.201304	0.761755
7	director 5	—	0.274506	1.036261
8	director 6	—	0.432347	1.468609
9	director 7	—	0.388884	1.857493
10	director 8	—	0.388884	2.246378
11	director 9	—	0.388884	2.635262
12	director 10	—	0.388884	3.024146
13	director 11	—	0.388884	3.413030
14	director 12	—	0.388884	3.801914
15	director 13	—	0.388884	4.191789
16	director 14	—	0.388884	4.579683

iterating the K2RIW antenna

The original thirteen-element K2RIW Yagi was just over 3.4 wavelengths long, and was probably designed for use in multi-Yagi arrays; this would explain the relatively short boomlength. Extending this design to sixteen elements results in a Yagi still readily stacked, but with even more impressive results for the single antenna. Baselined in **table 2**, this antenna is just over 4.5 wavelengths long and fits easily on a 10-foot boom. All elements are 0.125 inch in diameter, and the design frequency is 432.0 MHz. As is necessary for successful

table 3. A comparison of optimized 432 MHz K2RIW Yagis for each of six different director tapering approaches.

taper	optimized performance parameter	parasitic element lengths (inches) reflector	director 1	gain (dBi)	F/B (dB)
0.000000	gain	13.3125	11.5000	16.741	17.253
	F/B	13.4375	11.2500	16.372	32.666
0.015625	gain	13.3125	11.6250	16.737	17.535
	F/B	13.5000	11.3750	16.380	27.966
0.031250	gain	13.3125	11.7500	16.713	17.447
	F/B	13.5000	11.5000	16.400	25.026
0.046875	gain	13.3125	11.8750	16.696	16.971
	F/B	13.5000	11.6250	16.406	22.900
0.062500	gain	13.2500	11.9375	16.638	18.085
	F/B	13.6250	11.6875	16.211	21.372
0.078125	gain	13.1875	12.0625	16.595	17.428
	F/B	13.6250	11.8125	16.202	20.153

computer iterations, the driven element is specified at its R + j0 physical length.

Table 3 summarizes the results of six different pairs of K2RIW Yagi design iterations. As even the slightest element taper (increments of 0.015625 inch) results in degraded performance, full details are provided for only the zero taper antenna. Interestingly enough, Knadle specified a zero taper in the original design. Parasitic element lengths for optimizing a sixteen-element antenna differ from those that optimize a thirteen-element antenna. (More information on this phenomenon is contained in a later article in this series that addresses other Yagi optimization techniques.)

Table 4 presents the gain optimizing iteration that produced 16.741 dBi, and **table 5** presents the F/B optimizing iteration that produced 32.666 dB. **Tables 6** and **7** present these antennas'respective performance over the 8-MHz bandwidth. The gain optimized Yagi effectively illustrates the relatively broadbanded response of even long Yagis at 432 MHz. The F/B optimized antenna provides an excellent F/B across this same bandwidth. This optimized F/B is not a single frequency F/B resulting from sharp vectorial cancellation, but is the high level of F/B that comes naturally with long Yagis that are well designed. **Figures 2** and **3** present these antennas' respective E-plane plots. The gain optimized antenna has the sharper main lobe, but the F/B optimized antenna has a deeper first null. Knadle seems to have preferred a gain optimized antenna, but depending on the chosen antenna's designated use, either of these optimized K2RIW designs result in first-rate antennas. His original thir-

table 4. Optimized gain iteration for the zero taper K2RIW antenna with a reflector length of 13.3125 inches.

director 1 (inches)	gain (dBi)	F/B (dB)
11.0000	15.754	20.051
11.0625	15.918	21.305
11.1250	16.081	23.071
11.1875	16.241	25.658
11.2500	16.394	29.312
11.3125	16.533	30.406
11.3750	16.648	25.467
11.4375	16.724	20.842
11.5000	16.741	17.253
11.5625	16.676	14.346
11.6250	16.504	11.908
11.6875	16.214	9.836
11.7500	16.004	8.469
11.8125	15.336	6.733
11.8750	14.848	5.873
11.9375	14.075	8.464
12.0000	13.632	8.218

table 5. Optimized F/B iteration for the zero taper K2RIW antenna with a reflector length of 13.4375 inches.

director 1 (inches)	gain (dBi)	F/B (dB)
11.0000	15.727	21.757
11.0625	15.811	22.437
11.1250	16.058	25.663
11.1875	16.219	29.167
11.2500	16.372	32.666
11.3125	16.510	28.755
11.3750	16.621	23.484
11.4375	16.690	19.491
11.5000	16.696	16.335
11.5625	16.614	13.713
11.6250	16.424	11.474
11.6875	16.116	9.549
11.7500	15.704	7.929
11.8125	15.232	6.664
11.8750	14.766	5.902
11.9375	14.339	6.060
12.0000	13.616	8.739

teen-element Yagi provided a single frequency type of F/B when optimized for F/B.

the Tilton/Greenblum Yagi

The evolution of the Greenblum and Tilton/Greenblum designs was discussed in a previous article.[2] Tilton's 432 MHz Yagi was originally published as an eleven-element design.[6] For purposes of computer iteration and comparison with the Knadle design, a sixteen-element Yagi is presented. The additional five directors are spaced equally apart in accordance with

table 6. Frequency response parameters for the gain optimized zero taper K2RIW antenna.

frequency (MHz)	gain (dBi)	F/B (dB)
428.0	16.630	25.127
429.0	16.683	23.056
430.0	16.721	20.926
431.0	16.736	18.982
432.0	16.741	17.253
433.0	16.717	15.716
434.0	16.667	14.340
435.0	16.589	13.097
436.0	16.480	11.969

table 7. Frequency response parameters for the F/B optimized zero taper K2RIW antenna.

frequency (MHz)	gain (dBi)	F/B (dB)
428.0	16.122	23.635
429.0	16.191	25.519
430.0	16.256	27.874
431.0	16.316	30.675
432.0	16.372	32.666
433.0	16.422	31.198
434.0	16.466	27.976
435.0	16.501	25.028
436.0	16.526	22.584

table 8. Baselined Tilton/Greenblum Yagi at 432.0 MHz with fixed parasitic element spacings and parasitic element lengths supplied during each iteration.

element number	element name	length (inches)	element spacing (λ)	cumulative length (λ)
1	reflector	—	0.000000	0.000000
2	driven element	13.032811	0.137253	0.137253
3	director 1	—	0.137253	0.274506
4	director 2	—	0.173854	0.448361
5	director 3	—	0.237906	0.686266
6	director 4	—	0.265356	0.951622
7	director 5	—	0.274506	1.226129
8	director 6	—	0.347708	1.573837
9	director 7	—	0.347708	1.921545
10	director 8	—	0.347708	2.269254
11	director 9	—	0.347708	2.616962
12	director 10	—	0.347708	2.964670
13	director 11	—	0.347708	3.312378
14	director 12	—	0.347708	3.660086
15	director 13	—	0.347708	4.007795
16	director 14	—	0.347708	4.355503

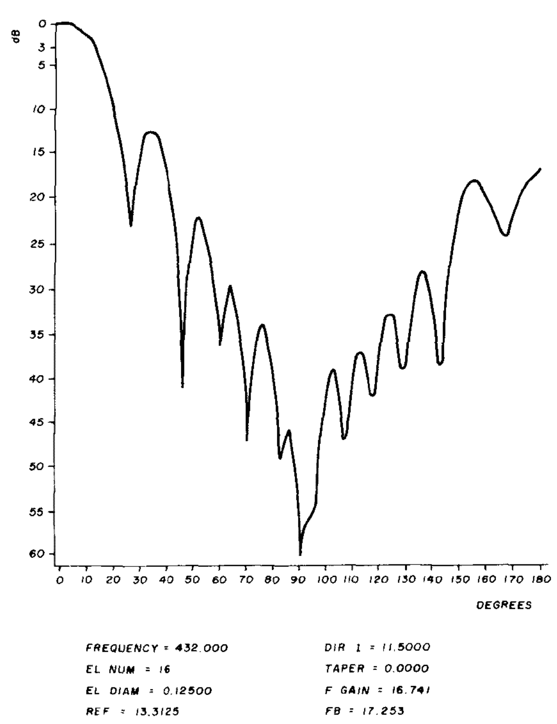

fig. 2. E-plane plot of the gain optimized K2RIW Yagi.

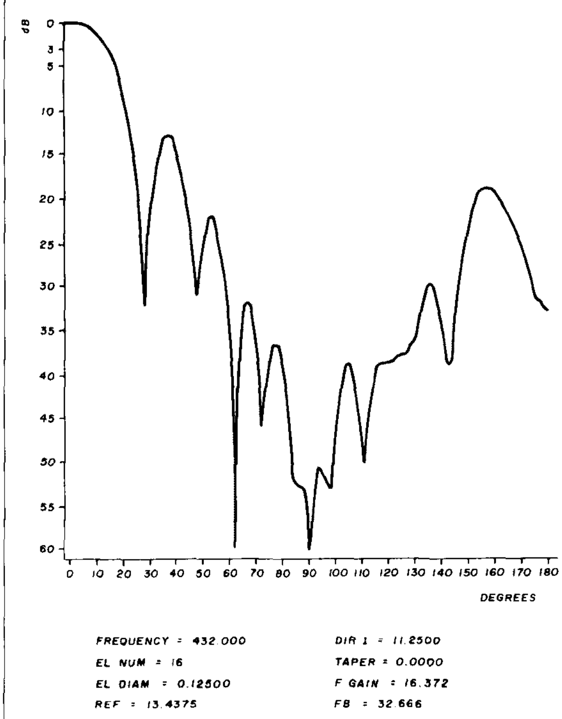

fig. 3. E-plane plot of the F/B optimized K2RIW Yagi.

table 9. A comparison of the optimized Tilton/Greenblum 432 MHz Yagis for each of thirteen director tapering approaches.

taper	optimized performance parameter	parasitic element lengths (inches) reflector	director 1	gain (dBi)	F/B (dB)
0.000000	gain	13.5000	11.6875	15.942	12.794
	F/B	13.1875	12.0625	14.303	38.225
0.015625	gain	13.5625	11.8750	16.103	14.277
	F/B	13.1875	12.1875	14.473	37.976
0.031250	gain	13.6250	12.0000	16.259	15.754
	F/B	13.2500	12.3125	14.753	35.519
0.046875	gain	13.6250	12.1250	16.390	17.088
	F/B	13.2500	12.3750	15.299	30.234
0.062500	gain	13.6250	12.1875	16.498	18.106
	F/B	13.2500	12.5000	15.187	42.848
0.078125	gain	13.6250	12.3125	16.574	20.076
	F/B	13.3125	12.5625	15.619	39.416
0.093750	gain	13.6875	12.3125	16.657	22.212
	F/B	13.3750	12.6250	15.907	37.921
0.109375	gain	13.6875	12.4375	16.634	24.416
	F/B	13.4375	12.6875	16.067	39.146
0.125000	gain	13.5625	12.5000	16.651	24.325
	F/B	13.4375	12.7500	16.099	42.827
0.140625	gain	13.5625	12.5625	16.583	25.312
	F/B	13.5625	12.7500	16.319	48.957
0.156250	gain	13.6250	12.6250	16.529	29.287
	F/B	13.7500	12.6875	16.457	46.691
0.171875	gain	13.6250	12.6875	16.448	30.353
	F/B	13.7500	12.6875	16.421	35.637
0.187500	gain	13.6250	12.6875	16.364	28.608
	F/B	13.8125	12.6875	16.319	31.650

table 10. Optimized gain iteration for a taper of 0.0625 with a reflector length of 26.375 inches.

director 1 (inches)	gain (dBi)	F/B (dB)
11.7500	15.903	17.498
11.8125	16.028	17.296
11.8750	16.143	17.153
11.9375	16.247	17.087
12.0000	16.337	17.122
12.0625	16.413	17.285
12.1250	16.469	17.605
12.1875	16.498	18.106
12.2500	16.485	18.778
12.3125	16.408	19.504
12.3750	16.235	19.941
12.4375	15.931	19.609
12.5000	15.484	18.497
12.5625	14.930	17.179
12.6250	14.359	15.962
12.6875	13.839	14.147
12.7500	13.355	11.391

table 11. Optimized F/B iteration for a taper of 0.0625 with a reflector length of 13.25 inches.

director 1 (inches)	gain (dBi)	F/B (dB)
11.7500	15.821	13.870
11.8125	15.936	13.701
11.8750	16.036	13.568
11.9375	16.119	13.495
12.0000	16.180	13.513
12.0625	16.215	13.660
12.1250	16.220	13.992
12.1875	13.665	22.300
12.2500	16.113	15.587
12.3125	15.986	17.210
12.3750	15.796	19.961
12.4375	15.535	25.365
12.5000	15.187	42.848
12.5625	14.753	24.093
12.6250	14.231	17.825
12.6875	13.664	13.723
12.7500	13.251	10.706

the Greenblum design data and *Tilton's* adaptation. This Yagi is constructed on a non-conductive boom and is baselined in **table 8**. Parasitic element diameter is 0.09375 inch, and the design frequency is 432.0 MHz. Tilton did not specify his own exact design frequency, but a careful reading of his article gives the *impression* that a slightly higher frequency was used. This would follow from the general band usage patterns of that era. As the matching of the transmission line and driven element is very critical on 432 MHz, Tilton's driven element assembly should be duplicated exactly. Any desired changes should be made in a sec-

table 12. Frequency response parameters for the gain optimized antenna with a 0.0625 taper.

frequency (MHz)	gain (dBi)	F/B (dB)
428.0	16.419	16.157
429.0	16.446	16.589
430.0	16.469	17.061
431.0	16.487	17.570
432.0	16.498	18.106
433.0	16.492	18.641
434.0	16.487	19.127
435.0	16.460	19.485
436.0	16.411	19.622

table 13. Frequency response parameters for the F/B optimized antenna with a 0.0625 taper.

frequency (MHz)	gain (dBi)	F/B (dB)
428.0	15.201	18.096
429.0	15.244	20.376
430.0	15.263	23.657
431.0	15.173	27.940
432.0	15.187	42.848
433.0	15.073	30.226
434.0	14.908	24.244
435.0	14.700	20.882
436.0	14.469	18.625

ond Yagi so as to allow comparisons with an established standard.

Table 9 summarizes the results of iterating thirteen pairs of Tilton/Greenblum 432 MHz Yagis. Performance trends are easily noted and there is no real need for additional iterations. As it is highly unlikely that many VHF/UHF operators could reliably measure or build to tolerances of 0.015625 inch, the Yagis of real interest are those at intervals of 0.0625 inch. These four Yagis also represent peaks or near-peaks in either gain or F/B. This same interval is also used for parasitic element lengths. Furthermore, the zero taper Yagis can be readily seen to be the poorest performers of these four pairs of antennas. As such they are excluded from the detailed examinations that follow for the tapering approaches of 0.0625, 0.125, and 0.1875 inch. The tables that follow refer to these Yagis.

taper = 0.0625

This is the only Tilton/Greenblum 432 MHz Yagi that was actually presented by Tilton. The two Yagis that follow (0.125 and 0.1875 taper) are the products of computer iteration. **Table 10** presents the gain optimizing iteration that resulted in 16.498 dBi of gain, and **table 11** presents the F/B optimizing iteration and its calculated result of 42.848 dB. Nearly 1.3 dB of gain and over 24 dB of F/B separate these two 0.0625 taper Yagis. **Tables 12** and **13** present these antennas' respective calculated performance over the specified bandwidth. Both antennas show peaks in their calculated performance parameters at 432.0 MHz. The F/B optimized value is clearly the result of single frequency vectorial cancellation. When coupled with a weak signal bandwidth that is relatively narrow in terms of the design frequency's wavelength, this F/B value will not deteriorate as rapidly as it would at a lower frequency. Extensive QSYs will probably result in noticeable decreases in rearward attenuation. **Figures 4** and **5** present E-plane plots of these antennas. The gain optimized antenna has a significantly sharper and deeper main lobe as well as a measurably higher degree of unwanted signal attenuation between 100 and 160 degrees. The F/B optimized antenna provides the expected outstanding level of unwanted signal attenuation from 160 to 180 degrees.

table 14. Optimized gain iteration for a taper of 0.125 with a reflector length of 13.5625 inches.

director 1 (inches)	gain (dBi)	F/B (dB)
12.0000	15.826	20.040
12.0625	15.968	20.006
12.1250	16.103	20.023
12.1875	16.229	20.117
12.2500	16.345	20.324
12.3125	16.433	20.524
12.3750	16.532	21.296
12.4375	16.592	22.247
12.5000	16.651	24.325
12.5625	16.601	26.163
12.6250	16.519	30.366
12.6875	16.359	35.590
12.7500	16.119	29.599
12.8125	15.825	24.253
12.8750	15.543	20.913
12.9375	15.355	18.600
13.0000	15.271	16.726

table 15. Optimized F/B iteration for a taper of 0.125 with a reflector length of 13.4375 inches.

director 1 (inches)	gain (dBi)	F/B (dB)
12.0000	15.826	18.494
12.0625	15.962	18.442
12.1250	16.090	18.422
12.1875	16.210	18.456
12.2500	16.318	18.573
12.3125	16.411	18.812
12.3750	16.486	19.229
12.4375	16.538	19.905
12.5000	16.558	20.971
12.5625	16.537	22.664
12.6250	16.460	25.498
12.6875	16.315	31.062
12.7500	16.099	42.827
12.8125	15.832	28.671
12.8750	15.567	22.795
12.9375	15.373	19.036
13.0000	15.262	16.159

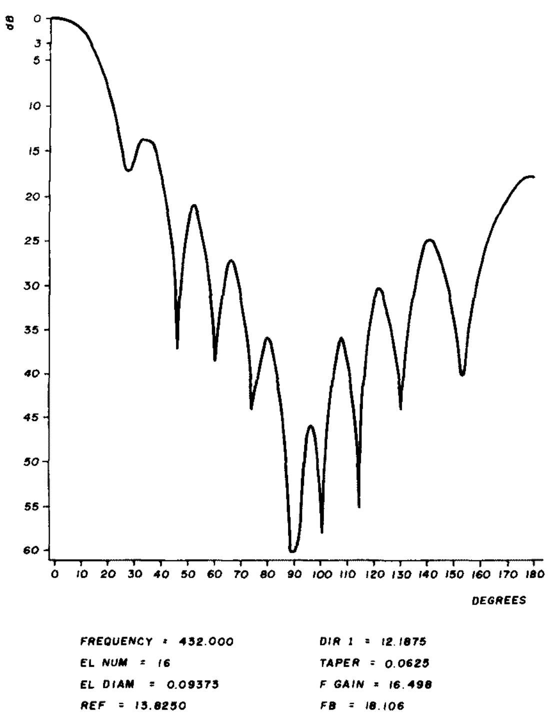

fig. 4. E-plane plot of the 0.0625 taper gain optimized Tilton Yagi.

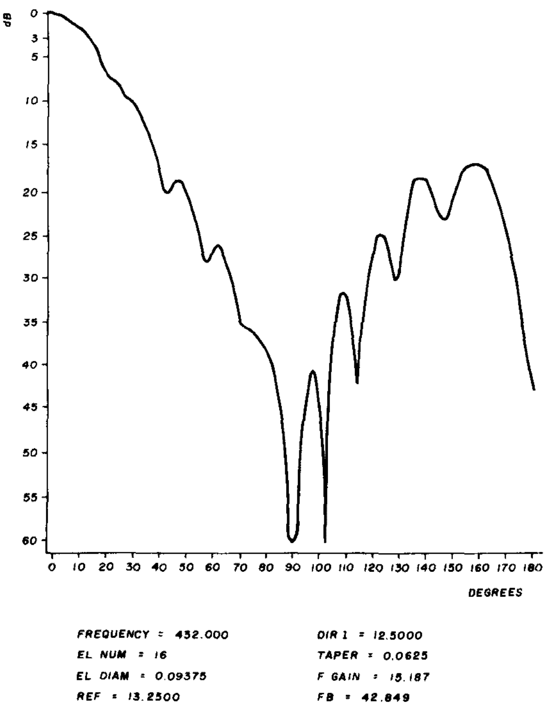

fig. 5. E-plane plot of 0.0625 taper F/B optimized Tilton Yagi.

taper = 0.125

This antenna exists solely as a product of computer iterations. In comparison with the 0.0625 tapered antenna design presented by Tilton and optimized through computer iteration, the optimized antennas with a 0.0125 taper require shorter reflectors and longer directors. **Table 14** presents the gain optimizing iteration that resulted in 16.651 dBi of gain, and **table 15** presents the F/B optimizing run with a calculated F/B of 42.825 dB. Over 0.5 dB of gain and 18 dB of F/B

table 16. Frequency response parameters for the gain optimized antenna with a 0.125 taper.

frequency (MHz)	gain (dBi)	F/B (dB)
428.0	16.498	19.253
429.0	16.542	20.106
430.0	16.578	21.101
431.0	16.604	22.290
432.0	16.651	24.325
433.0	16.622	25.548
434.0	16.608	27.807
435.0	16.575	30.379
436.0	16.520	31.856

table 17. Frequency response parameters for the F/B optimized antenna with a 0.125 taper.

frequency (MHz)	gain (dBi)	F/B (dB)
428.0	16.249	21.400
429.0	16.253	23.727
430.0	16.231	27.083
431.0	16.180	32.775
432.0	16.099	42.827
433.0	15.990	32.371
434.0	15.861	26.835
435.0	15.722	23.582
436.0	15.587	21.398

table 18. Optimized gain iteration for a taper of 0.1875 with a reflector length of 13.625 inches.

director 1 (inches)	gain (dBi)	F/B (dB)
12.2500	15.690	23.557
12.3125	15.830	23.848
12.3750	15.962	24.218
12.4375	16.083	24.698
12.5000	16.190	25.333
12.5625	16.277	26.172
12.6250	16.337	27.266
12.6875	16.364	28.608
12.7500	16.348	29.937
12.8125	16.281	30.455
12.8750	16.167	29.535
12.9375	16.023	27.980
13.0000	15.897	26.706
13.0625	15.845	25.256
13.1250	15.855	21.841
13.1875	15.733	17.729

separate these two antennas. **Tables 16** and **17** present these antennas' respective calculated performance over the relevant bandwidth. Both antennas' optimized parameters show peak values at 432.0 MHz. The F/B optimized antenna's F/B value is clearly a single frequency peak. However, as is true of the 0.0625 tapered F/B optimized antenna, the relative broad frequency responses of Yagis at 432 MHz is coupled with a relatively small bandwidth for the weak signal area. Consequently it is very likely that high F/B values, though not the calculated peak values, will be recognized in practice. **Figures 6** and **7** present these antennas' respective E-plane plots. The gain optimized antenna has the more sharply defined main lobe and generally deeper nulls throughout the pattern. Signal attenuation between 150 and 170 degrees is also greater than for the F/B optimized antenna. It is in the last 10 degrees that the F/B antenna provides a noticeable difference in performance. To some degree this F/B antenna typifies some of the exceptions to the use of F/B as an antenna performance parameter that were detailed by Lawson.[7]

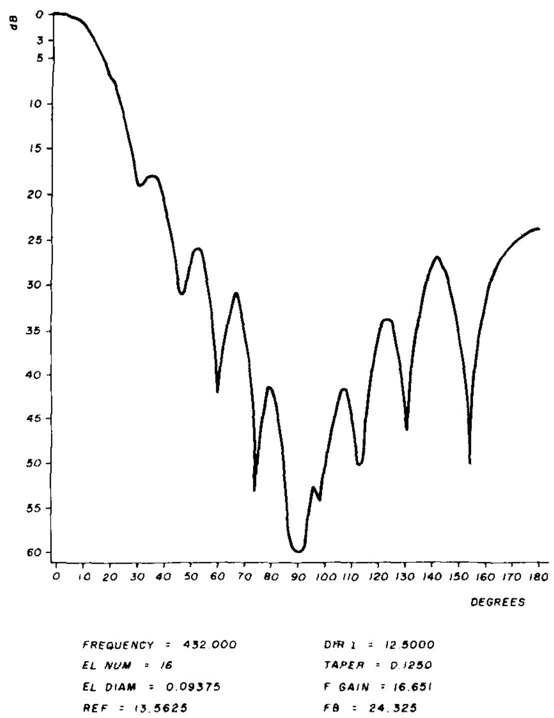

fig. 6. E-plane plot of 0.125 taper gain optimized Tilton Yagi.

taper = 0.1875

As is the case for the 0.125 tapered antennas, this pair of Yagis exists solely as the result of computer iterations. Compared to the 0.125 tapered antennas, longer reflectors and longer directors are used. **Table 18** presents the gain optimizing iteration that resulted in 16.364 dBi of gain, and **table 19** presents the F/B optimizing iteration with a calculated F/B of 31.650 dB. Less than 0.5 dB of gain and just over 3 dB of F/B separate these antennas. **Tables 20** and **21** present these antennas' respective calculated performance over the relevant bandwidth. Both antennas show peaks at 432 MHz for their respectively optimized parameters. The F/B values shown in **table 21** are not the single frequency F/B ratios of the two prior F/B optimized antennas. This F/B comes naturally from the antenna's design and is available over a 3 MHz bandwidth. **Figures 8** and **9** present these antennas' nearly identical E-plane plots, respectively. There are no distinguishing differences of any major consequence.

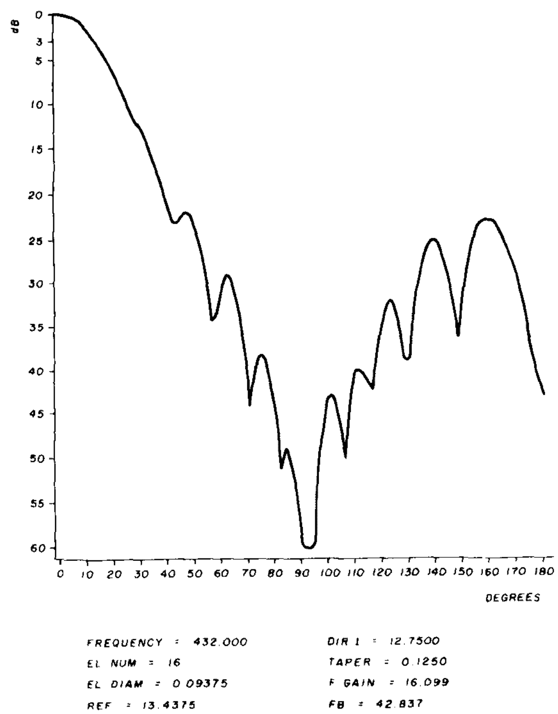

fig. 7. E-plane plot of 0.125 taper F/B optimized Tilton Yagi.

comparative results of computer iteration

These iterated comparisons between the single Knadle and multiple Tilton designs indicate there are no really significant differences in calculated forward gain. Any choice between these two design approaches, or from within the Tilton design approach would have to be based on a given antenna's calculated radiation pattern.

With the exception of the 0.1875 tapered Yagi, the F/B optimized antenna in each Tilton pair has the less clearly defined main lobe. The opposite was true of

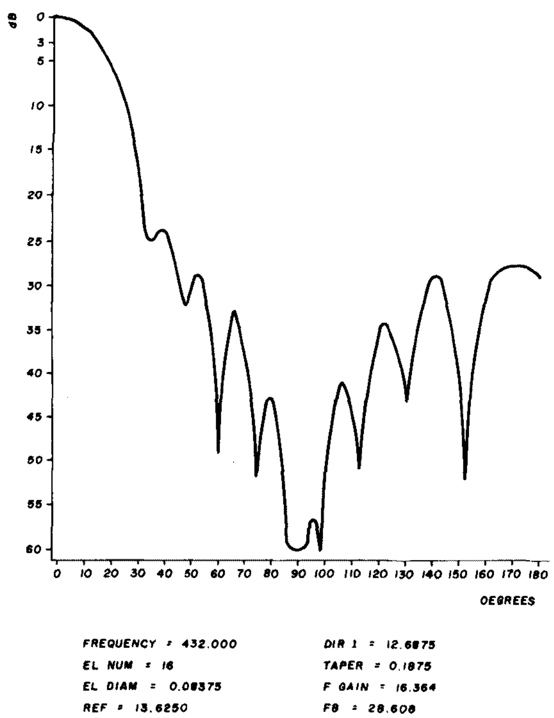

fig. 8. E-plane plot of 0.1875 taper gain optimized Tilton Yagi.

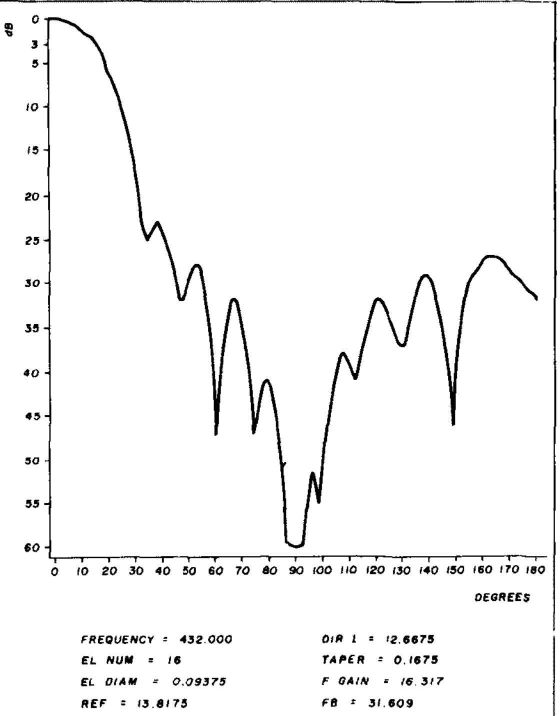

fig. 9. E-plane plot of 0.1875 taper F/B optimized Tilton Yagi.

the single Knadle Yagi. However, the first minor lobe on the Knadle Yagis tended to be of greater amplitude than the same lobe on comparable Tilton Yagis.

The need for an exceptionally high F/B on 432 MHz is open to question. F/B is a moot point for EME, and for terrestrial paths there simply is not the level of QRM comparable to 40 meters. Given that the pattern-wide attenuation of unwanted signals calculated for the Tilton 0.0625 and 0.125 gain optimized Yagis exceeds that of the F/B optimized Yagi in these pairs, these

table 19. Optimized F/B iteration for a taper of 0.1875 with a reflector length of 13.8125 inches.

director 1 (inches)	gain (dBi)	F/B (dB)
12.2500	15.627	25.171
12.3125	15.774	25.658
12.3750	15.912	26.283
12.4375	16.040	27.094
12.5000	16.151	28.143
12.5625	16.239	29.453
12.6250	16.299	30.867
12.6875	16.319	31.650
12.7500	16.291	30.726
12.8125	16.211	28.534
12.8750	16.083	26.331
12.9375	15.936	24.760
13.0005	15.823	24.066
13.0625	15.796	23.808
13.1250	15.782	21.524
13.1875	15.506	17.627
13.2500	14.849	15.885

table 20. Frequency response parameters for the gain optimized antenna with a 0.1875 taper.

frequency (MHz)	gain (dBi)	F/B (dB)
428.0	16.248	23.482
429.0	16.311	24.298
430.0	16.340	25.563
431.0	16.358	27.004
432.0	16.364	28.608
433.0	16.357	30.223
434.0	16.335	31.363
435.0	16.299	31.314
436.0	16.406	25.784

table 21. Frequency response parameters for the F/B optimized antenna with a 0.1875 taper.

frequency (MHz)	gain (dBi)	F/B (dB)
428.0	16.290	27.628
429.0	16.312	28.969
430.0	16.326	30.323
431.0	16.328	31.389
432.0	16.319	31.650
433.0	16.297	30.838
434.0	16.260	29.321
435.0	16.210	27.614
436.0	16.146	25.995

gain optimized Yagis seem to be preferable. To a much lesser degree this is also true for the 0.1875 Tilton Yagi pair. In effect the real choice of antennas is narrowed to any of the gain optimized Tilton Yagis and either of the Knadle Yagis.

The minor lobe *structure* of the 0.125 and 0.1875 gain optimized Tilton Yagis is measurably sharper than either of the Knadle antennas. If a final choice of antennas were to be made, either of these Tilton designs would be appropriate, with the 0.125 design preferable. The selection of one antenna design from these many fine designs depends more on the station operator than any of the calculated differences in the stated selection criteria. Excellent results can be expected from either of the two preferable Tilton designs or from either of the Knadle Yagis. The differences in forward gain are more a function of the Knadle Yagi's longer boomlength than any verifiable design superiority. In fact, the additional 0.22 wavelength advantage should have given the Knadle Yagi a larger margin by at least another 0.1 dB.

Comparisons among these Tilton and Knadle Yagis and the NBS 4.2 wavelength Yagi are also interesting. Though the former Yagis are only marginally longer, they provide higher calculated forward gains. The NBS Yagi has a calculated forward gain of 15.71 dBi. The 0.0625, 0.125, and 0.1875 Tilton gain optimized Yagis were calculated to have 16.498, 16.651, and 16.364 dBi of gain, respectively. The gain and F/B optimized Knadle Yagis had calculated gain figures of 16.741 and 16.372 dBi, respectively. The additional gain calculated for any of these five Yagis is in excess of what the comparative differences in boomlength would explain. An additional director and the use of varying parasitic element spacing are the main reasons for these differences. The NBS antenna's first minor lobe is of greater amplitude than these 0.0625 and 0.125 Tilton Yagis. The NBS antenna also has slightly less F/B than these 0.125 and 0.1875 Tilton Yagis.

Next month's article will present a single pair of 50 MHz Yagis with highly desirable performance parameters. They represent an interesting exception to a very popular design approach.

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ham radio

ham radio TECHNIQUES

Bill Orr W6SAI

160 redux

Fifty years ago 160 meters was the beginner's phone band. Use of the 80- and 20-meter phone bands was restricted to the Class A (Advanced) licensee, 15 meters was not an Amateur band, 10 meters was sparsely occupied by experimenters and the 40-meter phone band did not exist.

During weekdays, until about 5 PM local time, 160 phone was a cross-section of juvenile America. High school lads, with plenty of enthusiasm and little money, were on the air with 20 or 30-watt phone rigs. In the evening hours, they were obliterated by older, more affluent Amateurs who boasted powerful rigs of 50 to 150 watts. All in all, it was a lot of fun.

There was plenty of DX excitement for the young operators. Many a tired teen-ager stumbled into class in the morning after an all-night DX session. On the east coast, the mark of a true DXer was that he had been able to hear W6JYH in Los Angeles on phone. (*Working* him was out of the question!) On the west coast, the test was to hear W1DJL in Massachusetts. And everybody envied the booming 160 phone signal of W4CPG in New Port Richey, Florida.

With the advent of World War II and the invasion of LORAN, the old 160-meter band remained practically deserted except for a few die-hards — until recently. Now, with the band appearing on the bandswitch of most new exciters, and LORAN banished to the low frequencies, interest in "top band" has exploded.

I listened in during the spring CW contest. The band was jumping with CW signals from all over the United States and Canada, crowding the

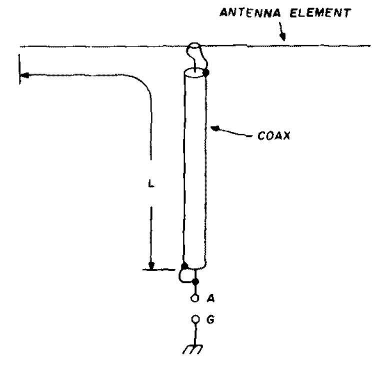


fig. 1. The active element of an HF antenna plus the feedline can be used as a T-Marconi antenna on 160 meters. One-half the element length plus the feedline length (L) is the length of the 160-meter antenna. Note that the coax conductors are jumpered at the feedpoint (A-G).

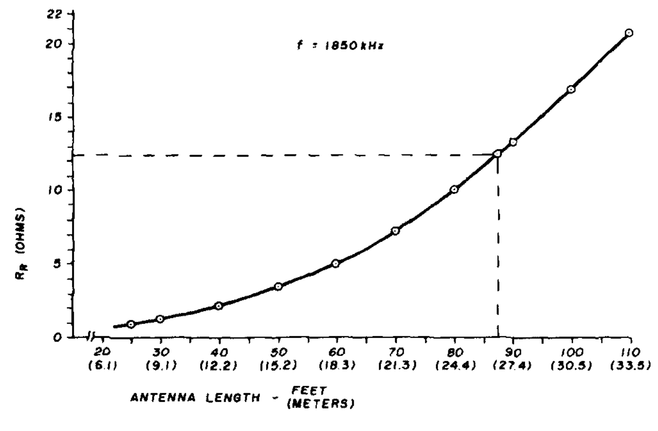


fig. 2. Approximation of feedpoint resistance (R_R) for a short 160-meter Marconi antenna.*

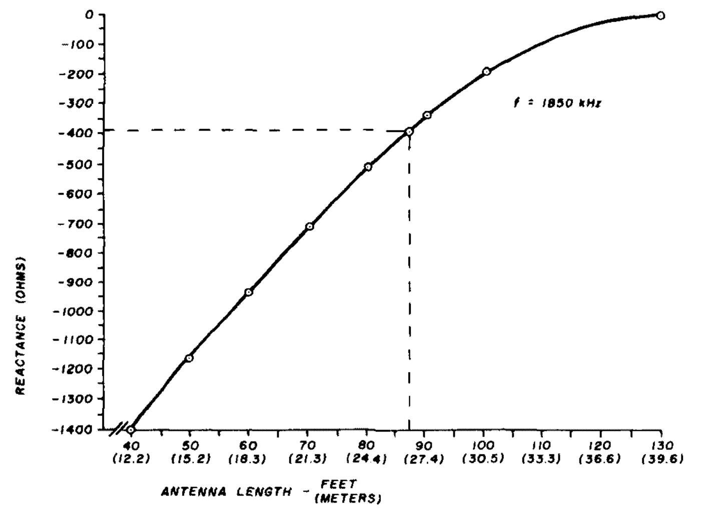


fig. 3. Approximation of the feedpoint reactance for a short 160-meter Marconi antenna.*

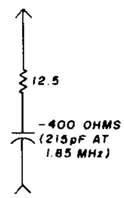


fig. 4. Equivalent circuit of my 160-meter Marconi antenna.

band from 1800 to about 1850 kHz. And there was plenty of SSB activity from 1850 to 1950 kHz, with some contented AM operation at the top end of the band.

Yes, 160 had come back to life and I had to get on. All I needed was an antenna.

let's get on 160 meters

One possible choice is to use the high frequency antenna (dipole, Yagi, Quad, or groundplane) as a T-type Marconi antenna, working the antenna-plus-feedline against ground (**fig. 1**).

This can be accomplished by shorting the inner and outer conductors on the coaxial line at the station and using a good earth ground connection as the return. If half of your driven element plus the feedline length approaches a quarter wave (about 126 feet) on the top band, you're in good shape; a simple antenna tuner will get you on 160 meters quickly. Just make sure your coaxial line is clear of the ground and not bundled up with rotator cables and such as it functions on 160 meters as part of the radiating system.

If your regular antenna tuner covers 160 meters, it may work with overall antenna lengths between 115 and 150 feet. But what do you do if your antenna-plus-feedline length is shorter than this? A different approach to the problem is required.

very short antennas for 160 meters

I measured my feedline length, plus half the driven element length of my tribander and the total length came out to be only 87 feet 6 inches. Could I work this shortened antenna system on 160 meters?

First, I had to determine if the physical length (L) was the same, or close to, the electrical length. I connected the T-antenna to my ground connection via a small 4-turn coil and used a dip oscillator to determine the natural quarter-wave resonance of the system. It was 2.73 MHz. That agreed closely with the theoretical resonance value, which is:

$$f(MHz) = \frac{234}{length\ (feet)} = \frac{234}{87.5} = 2.67\ MHz$$

Now that I knew the electrical and physical length of my proposed 160-meter antenna were very close, I could determine the radiation (feedpoint) resistance of the antenna, assuming I had a perfect ground. **Figure 2** shows that value to be close to 12.5 ohms. (See note, **fig. 2**)[1]

For my ground I ran a short, heavy conductor to the copper plumbing system of the house and grounded the pipes via two eight-foot long ground rods driven into the soil at each end of the house. (Unfortunately, the lawn sprinkling system buried in the front yard was constructed of PVC plastic pipe, otherwise it would have been added to the ground system too.)

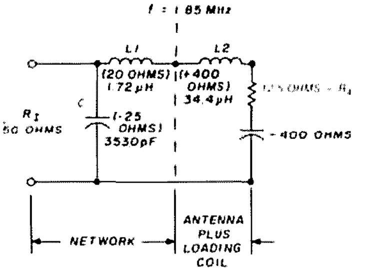
I decided to use a simple L-matching network between my transceiver and the 160 meter T-antenna and concluded it would be best to design this for the lowest estimated feedpoint resistance of my antenna because it's much easier for the network to match a load higher in resistance than the design value than lower in value. Accordingly the worst-case feedpoint resistance was chosen to be 10 ohms. Because I didn't know the ground resistance, and didn't have the time or inclination to measure it, I ignored it for the time being.

antenna reactance

I wasn't out of the woods yet. An antenna shorter than an electrical quarter wavelength exhibits negative (capacitive) reactance at the feedpoint, with the exact value depending upon the antenna length and the length/diameter ratio of the conductor. For a thin wire, the chart in **fig. 3** provides an approximation. (See **fig. 3** footnote.)[1]

My particular antenna would have a negative reactance of about −400 ohms. A schematic representation of what the antenna is, as far as the transceiver and matching network is shown

*Derived from data presented in reference 1.



$$C = \frac{\sqrt{R_A(R_I - R_A)}}{2\pi f \times R_A R_I} \times 10^6\ pF = 3530$$

$$L_1 = \frac{\sqrt{R_A(R_I - R_A)}}{2\pi f} = \mu H = 1.72$$

$$L_{TOTAL} = L_1 + L_2 = 34.4 + 1.72 = 36.12\ \mu H$$

Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms k = 1,000 M = 1,000,000

fig. 5. Pi-L network (L1C) plus loading coil (L2) are used for a short Marconi antenna. General design formulas are shown.

in fig. 4. Expressed as a complex number, the load to be matched is about 12.5-j400. The resistive portion of the load can be estimated more accurately than the reactive portion, which varies widely with antenna capacitance to ground and nearby objects.

In any event, the requirement is a 4:1 impedance transformation to 50 ohms plus cancellation of the −400 ohms of capacitive reactance. The cancellation can be accomplished by adding +400 ohms of inductive reactance between the antenna and the matching device. The required reactance is termed a *loading coil.*

the matching network

The equivalent antenna circuit shown in fig. 4 has to be accurately matched to 50 ohms in order for modern transceivers to work properly on the top band. Because of limited adjustment range, most transceivers are not very "forgiving" with regard to a mismatched antenna load, particularly on the 160 meter range.

A matching network to transform approximately 12 ohms to 50 ohms and cancel out the negative reactance exhibited by the antenna is shown in fig. 5. The circuit is an L-network, with extra series inductance added to cancel the capacitive effect of the short antenna. In practical terms it boils down to lots of shunt capacitance (C) plus a modest amount of series inductance (L1) in the network, plus the large extra loading inductance (L2) required as a loading coil. The two coils can be combined into one, as shown in the assembly of fig. 6.

Surplus mica capacitors are used, plus a large variable capacitor. A multigang "broadcast" capacitor with the sections tied in parallel can be substituted for the single section capacitor used in this unit.

it works — almost

In real life, the required inductance turned out to be less than the calculated amount, indicating that the antenna had less capacitive reactance than determined by the design chart. The antenna was measured to actually be 11.8 ohms in series with −211 ohms (407 pF).

After all was tuned up and the SWR reading between the tuner and the transceiver was unity, the last step was to decouple the house wiring from the antenna system. As soon as I got on the air, the lights in the family room blinked on and off in step with my keying.

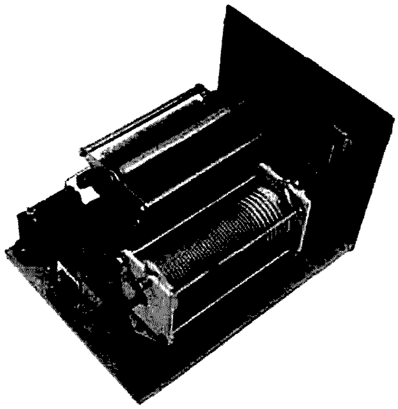


fig. 6. The 160-meter antenna tuner is built up on a plywood base and Masonite panel. The variable capacitor is a surplus Cardwell 2000 pF unit — probably unobtainable! Connected in parallel with the variable capacitor is a 0.002 μF transmitting-type mica capacitor and across it, a 250 pF mica unit.

It seemed as if a goodly proportion of my radiated power was being absorbed by the nearby house wiring. Since I couldn't move the antenna, or the house, the only thing left to do was to decouple the wiring from the RF field as much as possible. This was accomplished by taking a few AC plugs and mounting 0.01 μF, 1.6 kV ceramic capacitors across the pins.* The plugs were then inserted into wall outlets around the house at random until no lights came on when I was on the air and turning lights off and on did not affect the SWR reading. No kidding, when the house wiring was "hot," switching lights on and off *did* change the 160-meter antenna tuning and the SWR on the transmission line! (Live and learn.)

This all came about because the exposed house wiring was in close proximity to the antenna. It can happen on other bands, too. The best situation is to live in a home where the wiring is fed through metal conduit.

the results

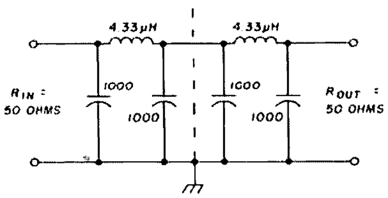
After all of this horsing around (which took only about a day of spare time) I settled down to see what I could work with this puny antenna. While I was not the biggest kid on the block, I could do well within 500 miles and could get reasonable reports out to the Midwest. I capped my DX activities one weekend by working two stations in New England! Granted, I got only a 339 report from each of them, but the contacts were genuine.

a 160-meter harmonic filter

Part 97.73 of the FCC rules specifies that spurious signals generated by the transmitter must be attenuated 40 dB below the output power level at frequencies between 1 and 5 MHz. Most Amateur exciters can meet this require-

*The "garden variety" 0.01 μF, 600-volt disc ceramic capacitor is not rated for operation at 115 VAC. Various manufacturers market special capacitors for this purpose that are rated for continuous operation at 125 VAC and 1400 to 1600 VDC. (The high DC rating is required because of high-voltage transients that often appear on the power line.) Three well-known brands of capacitors are Aerovox type AC-7, Centralab type CI-103, and Sprague type 125L-S10.

ment with room to spare, but this doesn't help the nearby Amateur working DX on 80 meters when he or she picks up your weak second harmonic atop the weaker, more desirable signal. A few extra dB of harmonic attenuation will solve this problem.



Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

fig. 7. Harmonic filter for 160-meters. Inductors are 1-inch diameter, 24 turns No. 18, 8 turns per inch (24-1/2 turns).

Shown in **fig. 7** is a simple 5-element Chebyshev lowpass filter that will attenuate the second and higher harmonics of a 160-meter signal about 24 dB, provided the SWR on the transmission line is near unity. The filter is placed immediately after the transceiver and the SWR meter and before the tuner.

Construction of the filter is simple. A two-compartment box is soldered up out of circuit board and a filter section is placed in each compartment. The size of the box depends upon the size of the components. For powers below 150 watts, postage-stamp size mica capacitors can be used, together with miniature air-wound coils. Matching coaxial receptacles are placed on the ends of the box.

The cutoff frequency of this filter is just above the top edge of the 160-meter band, so be sure to remove it from your antenna circuit when you operate on the higher bands, or you'll have fireworks.

in summary, then . . .

In order to use your HF beam, dipole, or whatever, with your feedline as a Marconi antenna, you'll have to determine the electrical length of the installation for 160-meter operation. This measurement is from one tip of the antenna element to the station. You can readily see that the coaxial line comprises the greater portion of the antenna. If the line runs close to the ground, performance on the top band won't be very good, because your makeshift antenna will be low. If the line runs 10 to 20 feet above the ground, you are in luck. Don't worry about twists or turns in the line; the antenna tuner will take care of that.

Watch out for coupling between your antenna and the house wiring or other nearby antennas. Some time ago, during an earlier attempt to get on 160 meters, I loaded up a random, 100-foot wire I'd used for general reception with my vintage National HRO receiver. During the tuning process I noticed smoke coming out from underneath by beloved receiver. I turned everything off quickly and soon discovered that the 15-foot long lead-in to the HRO antenna post was draped near the wire serving as part of the 160-meter antenna. Enough RF pickup existed to fry the receiver antenna coil to a crisp, even though no physical connection existed between the transceiver and the receiver.

The point is that a large portion of the 160-meter antenna was *in* the house, or near to it, and a strong RF field existed where one had never existed before, because the usual high-frequency antenna is not in close proximity to the house. *Any* unbalanced Marconi-type antenna working against ground is capable of pumping large portions of RF into the house wiring, if you don't take precautions to prevent it!

So there you have it. 160-meter operation is possible even with a small antenna *if* you have a reasonably good ground connection and the right antenna tuner. Keep clear of your house wiring, get on the band, and I'll see you on the low end!

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ham radio

ground rod resistance

Exploring ground, ground rod, and antenna relationships

An approximate equation has been derived for determining the resistance of a ground rod under a quarter wave vertical antenna. The physical factors involved are the antenna height and ground rod length and diameter, combined with the resistive and dielectric properties of the ground under the antenna. The derivation is based on choosing earth current paths which are both realistic and expressible with simple equations. With the equation it can then be demonstrated that in the regions where the current path is less certain, the power loss and hence the resistance contributions are minor.

The electrical characteristics of the ground are that of a "quasi-conductor" because of its dielectric property.[1] At 160 meters, it behaves more nearly as a resistive body, but at 10 meters, the dielectric currents become predominant. The effect is to markedly *reduce* the "ground resistance" at the higher frequencies.

The "constriction resistance" to the radial current as it leaves the ground rod can be calculated. As an example, at 80 meters, over one-third of the ground rod resistance is located within 6 inches (12.7 cm) of a 1-inch (2.54 cm) diameter ground rod. It confirms that the injection of salt brine into this very small region is of great benefit in reducing the effective ground resistance.[2]

The equation will answer such questions as what would be the effect of changes to the ground system. Such changes might be the use of longer or greater diameter ground rods, or the addition of more, well separated ground rods.

A ground rod can be considered as an inverted antenna in a lossy dielectric. At 10, 15, and even 20 meters, adverse resonances can occur, adding to the ground resistance. The approximate equation derived in this article does not include this effect, although the conditions for its presence are discussed in an appendix.

approximate equivalent circuit for ground currents

When a vertical quarter wave antenna is used with a *ground rod*, the current which enters the ground rod leaves it along its length, just as the antenna current leaves along *its* length. **Figure 1** shows the antenna system considered here, sketching the vertical plane through the antenna. Dimensions h, g, and d indicate *antenna* height, ground rod length, and ground rod diameter, respectively, and will appear in the ground rod resistance equation. The assumed current paths are discussed later.

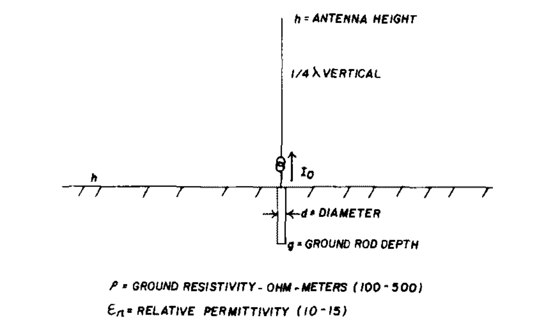

fig. 1. Antenna and ground system.

Evaluation of the ground currents which flow from the ground rod requires a knowledge of the electrical characteristics of the ground. Ground has both resistive and dielectric properties.[1,3] Resistivity is the reciprocal of conductivity, the latter being the characteristic usually tabulated. In this article, resistivity will be used in the ground rod equation. However, conductance may be used in the discussion, when appropriate.

The normal ranges for resistivity and relative permittivity are given in **fig. 1**. The resistivity shows such great variability that at a particular site it probably should be measured. This can be done simply and accurately with a four-point probe.[4]

Fortunately the relative dielectric constant seems to be less subject to variation. A major contributor is the

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moisture content. A choice of 12 to 14 could suffice, and is adequate for the present article.

These two fundamental constants will now be related to approximate discrete equivalent circuit elements. A cubic meter of ground is shown in **fig. 2.** Two one-meter square conducting plates are placed on opposite sides. Such a meter cube has the property that when using MKS (meter-kilogram-second system of units — metric) units, the resistance between the two plates is numerically equal to the resistivity. Thus from a measurement or by estimation the parallel resistance can be determined for the approximate equivalent circuit shown below the cube in **fig. 2**

A representative relative dielectric constant, $\varepsilon_r = 13$, will usually be used for discussion, and preparation of figures in this article. The capacitance for this value is 115 pF.

The formula for the capacitance between two parallel plates is from reference 5. Because the cube obviously conducts DC, the approximate equivalent network capacitance must appear in parallel.

When RF current flows through the ground, both equivalent parallel elements conduct. At very low frequencies, almost all goes through the resistor. As the frequency increases, an increasing part passes through the lossless dielectric. Only current through the resistor causes power loss. As shown in **fig. 2**, the magnitude of the resistive part of the current is:

$$I_R = \frac{I}{\sqrt{1 + Q^2}}$$

where Q from **fig. 2** introduces the current division effect. The sum of all the power losses, P, determines the ground rod resistance through the relation:

$$R_g = \frac{P}{I^2} \qquad (1)$$

The ground behaves as an inductive impedance, rather than simply as a resistance.[6] (This is discussed in **appendix B**.) The inductive effect is caused by the lower velocity of wave transmission through the dielectric, compared to that through the resistance, acting like current lagging a voltage. *This inductive effect contributes to a vertical antenna being shorter than the length called for by the formula*; the negative phase angle of the antenna is used to neutralize the positive phase angle of the ground.

ground resistance equation

The approximate equation for the ground rod resistance is:

$$R_g = \frac{\varrho}{(1 + Q^2)} \cdot \frac{1}{2\pi g} \left[\left(\log_e \frac{2h}{d}\right) - 1\right] \text{ ohms} \qquad (2)$$

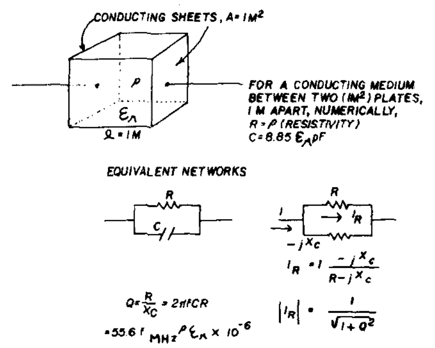

fig. 2. Ground electrical characteristics.

where h = antenna height, meters
g = ground rod length, meters
d = ground rod diameter, meters
ϱ = ground resistivity, ohm-meters
$Q = 55.6 f_{MHz}\, \varrho \varepsilon_r \times 10^{-6}$, from **fig. 2.**

The derivation of this equation is given in **appendix A.**

As an example, take a 20-meter high, quarter wave vertical for 80 meters, with an 8-foot ground rod 1 inch in diameter, and a ground resistivity of 200 ohm-meters. Using the equations in **fig. 2**, Q = 0.5. Then:

$$R_g = \frac{200}{(1 + 0.5^2)} \cdot \frac{1}{2\pi \cdot 2.44} \left[\left(\log_e \frac{2 \times 20}{0.025}\right) - 1\right]$$

$$= 66.5 \text{ ohms}$$

(equal to about that for eight radials.[9])

If the ground rod diameter were increased to 12 inches, the ground resistance would be 40.5 ohms, about 40 percent reduction. Thus, as mentioned earlier, over one-third of the ground loss occurs within 6 inches of a 1-inch rod. One half is inside the 1-foot radius.

It is convenient to consider this equation as the product of two factors, which have been plotted as curves in **figs. 3** and **4. Figure 3** represents how the ground resistivity, with its relative dielectric constant and the frequency, enter into the equation. **Figure 4** represents how the one-quarter wavelength vertical antenna height and the relatively shorter ground rod dimensions enter. The product of the two factors determined by these curves is the ground rod resistance in ohms.

equivalent series resistivity

The curves of **fig. 3**, one for each frequency band, are directly proportional to the antenna's ground resistance. The ground resistivity usually is reported as being in the range from 100 to 500 ohm-meters.[3]

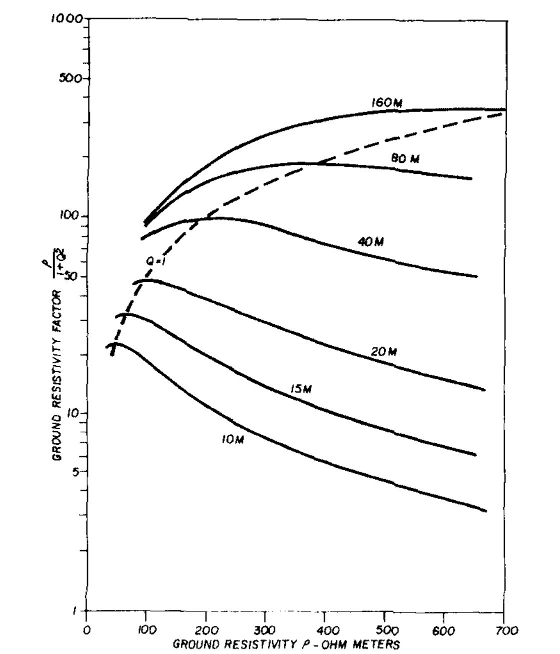

fig. 3. Ground equivalent series resistivity for $\epsilon_r = 13$ *(115 pF for 1M³).*

These curves can be considered as plots of the equivalent series resistance determined by a fixed capacitor shunted by a range of resistances. For each curve, the plot is for a shunting resistor increasing from less than the reactance of the capacitor, up to 700 ohms. Such a network has the property that the maximum equivalent series resistance occurs when the parallel resistance is numerically equal to the reactance, and for that condition, that is $Q = 1$, the series resistance maximum is exactly equal to half the parallel resistance.

The curves for 40 and 80 meters surprisingly show that because the maximum is centrally located, the series resistance changes less than 2 to 1 over the normal ground conductivity range.

The reactance of the parallel capacitor decreases inversely with increasing frequency. Thus the corresponding equivalent series resistivity peak decreases, with each curve becoming much lower overall. Two advantageous effects appear here. Not only is the peak less, *but because it moves outside the parallel* resistivity range normally encountered, the working equivalent series resistivity range is much lower than the peak. For example, at 10 meters, the effective resistivity is decreasing to below 10 ohm-meters, less than 1/20th of that for 80 meters. This is because an increasing portion of the ground current has become lossless dielectric current.

antenna system dimensions

The curves of **fig. 4** are drawn using the ratio of the one-quarter wavelength high antenna to the ground rod diameter as the x-axis. The height of the antenna is taken as representing the radius on the ground beyond which practically no ground currents flow. The curves are for different ground rod lengths. A possible adverse resonant effect not included in the approximate equation and which may be important at 10 and 15 meters, is considered in **appendix B.**

The chart is entered through the x-axis. The ground rod length is found next. From there across to the y-axis determines the dimension factor.

Repeating the 80-meter example calculated earlier, from **fig. 3**, the ground resistivity factor is 150. From **fig. 4**, for $g = 2.4$ *meters*, the dimension factor is about 0.42. The product is 63 ohms, duplicating the earlier result.

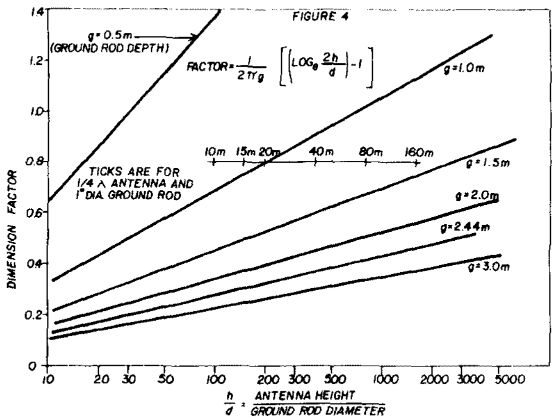

fig. 4. Ground rod resistance can be found using this chart in conjunction with ***eq. 2.***

For convenience, a bar is drawn with ticks shown for each Amateur HF band, locating the h/d for 1-inch diameter ground rods and quarter-wave verticals. For other diameters, start with a tick for 1 inch, then move right or left inversely as the ratio of the 1-inch diameter to the diameter wanted. For instance, for a 1/2-inch diameter ground rod, move right to twice the 1-inch h/d value. Thus at 20 meters, move from a scale of 200 to 400. Then locate g to find the dimension factor.

The curves clearly show that there is a large decrease in effective ground resistance as the frequency increases. The dimension factor from **fig. 4** is also lower for the higher frequencies. Approximate ground resistance measurements at 10, 15, and 20 meters to support the results of these calculations will be given later.

parallel ground rods

The parallel resistance of two ground rods may be estimated through a study of the current flow paths. When there are two ground rods, each receiving half of the total current, each acts as an "image" for the other. This effect causes the resistance of the two in parallel to be higher than just half of either alone.

The image repels any current flow across a perpendicular plane which bisects the line joining the two rods. The bisector thus acts as a virtual insulator. Visualization of the current flow paths from the two rods is helped by field maps shown in many texts.[8] They usually are drawn for static charges, but the field lines are like current flow lines in sheets.

Near each rod and inside a radius of about half the distance to the bisector, the effect of the other rod, while increasing, is small. Inside this region then, almost the full advantage of separate rods is realized. A resistance factor of 1/2 will be used.

For the next region, to a radius of the distance to the bisector, all currents including those which started out toward the bisector have completed most of their change in direction. The advantage of separate ground rods thus decreases in this region. A resistance factor of three quarters will be used.

Beyond this second region there is very little gain. To apply the above discussion an estimation involves first calculating the external resistance for circles centered around a single ground rod. This is done for diameters of 2 feet, 3 feet, and so on, using the formula. Such calculations for 80 meters are shown in **fig. 5.** Two cases have been chosen as examples, applying to 4-foot and 6-foot spacings.

For **Case 1**, the estimate of the resistance using two ground rods spaced 4-feet apart is shown as 48 ohms or a reduction compared to a single rod of 27 percent. For comparison, about 14 radials have a resistance of 48 ohms.[9] The calculation for the first foot assumed a 0.5 factor as complete independence, as discussed earlier. The second foot used a 0.75 factor for an intermediate effect, while outside that, negligible reduction compared to a single ground rod was used.

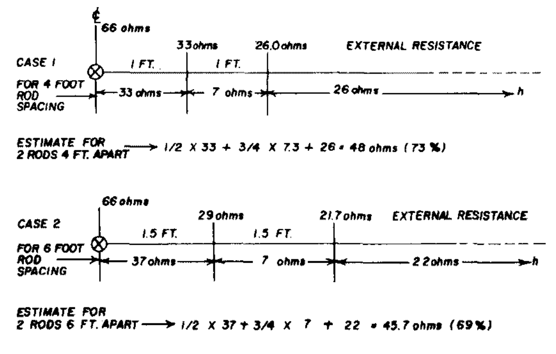

fig. 5. Estimation of resistance at 80 meters using two ground rods.

A very large number of rods in the circle could only approach 26 ohms. Thus two rods accomplish almost half of the possible reduction.

Case 2 is for rods spaced 6 feet apart. Calculating in the same manner there is a reduction to about 45 ohms, or a 31 percent decrease. The increased spacing from 4 to 6 feet thus gained another 2.3 ohms.

20 meter measurements

The striking information of **fig. 3** and **4** is the exhibition of the beneficial effect of the ground's dielectric property in reducing the effective ground resistance. With increasing frequency, the increasing Q rapidly reduces the resistance.

Elementary but restricted resistance measurements were made with a double-pointer SWR bridge. The restriction with this method is that the resistance can be determined only at a frequency where the SWR is a minimum.

Starting with a vertical wire having a 16 foot 9 inch formula length, the wire was shortened until the minimum was within the band (final length was 15 feet, 10 inches). Using two rods four feet long and four feet apart, measurements and calculations were as follows:

rods	SWR	R_g ohms	R_g (calc.)
1″ D	**1.43***	34.5	27
3/4″ D	1.33	28.0	30
parallel	**1.18***	19.0	20 (est.)

The addition of a shunt resistor* has reduced the SWR in the measurements printed in bold type, establishing that the load was greater than 50 ohms. The formula for such a minimum SWR, is:

$$R_g = [(50 \times SWR) - 37.5] \text{ ohms} \qquad (3)$$

These results agree to within a few ohms of the resistances calculated using the formula. The measurement formula for R_g may be optimistic in that the radiation resistance of the antenna probably decreases from 37.5 ohms as the wire is shortened for system resonance.

10 and 15 meter measurements

There was further confirmation at 10 and 15 meters when measuring with a shorter 2 foot, 3/4-inch diameter round rod:

band (meters)	formula height	tuned height	min. SWR	R_g ohms	calc. R_g ohms
15	11′ 0″	9′ 9″	1.49	37	33
10	8′ 4″	7′ 7″	1.14	19	17

Even with a 2 foot ground rod, at 10 meters (*10 meter* band) the resistance was about half that at 20 meters with a *4 foot* ground rod.

However, a problem was found when trying to measure using the 4-foot ground rod. As the antenna height was reduced, the SWR curve kept the same slope across the band, reached a lowest curve, and then increased again. There was no minimum in the band. Thus the SWR bridge method could not be used. The lowest SWR, at the band edge for both bands, was higher than the minimum found for 20 meters.

The lower end of the SWR curve was on the high frequency edge for 15 meters and on the low frequency edge for 10 meters. It may be that a resonance in the 4-foot ground rod, acting as an inverted antenna in a lossy dielectric prevents the lowest SWR curves having a minimum in the band. (Further discussion is in **appendix B**.) At 10 and 15 meters, for the local conductivity condition, one-quarter wavelengths are 2-1/2 and 3 feet respectively. Thus the 4-foot *ground* rods were above their series resonance. It may be that an odd number of underground quarter wave lengths is best.

conclusion

An approximate equation for the resistance of a ground rod has been derived when using a quarter wave vertical antenna. It relates the elements which affect the ground rod resistance and provides a means for estimating the result of changes in the ground system. Measurements confirm the importance of the relative dielectric constant of ground in reducing the effective ground resistance as the frequency increases.

How well the formula applies to trap verticals has not been established. Knowing the trap antenna resistance for the subtraction is another problem.

With the use of ground rods only there always is some power loss penalty. However, sometimes pavements, sidewalks, buildings, or someone else's property precludes other approaches. Even with such a restriction, it appears that a system equivalent in power loss to quite a few radials can be realized — but it is unlikely that the radiation angle will be as low.

appendix A

derivation of the approximate ground resistance equation

This equation applies to the antenna and ground rod system shown earlier in **fig. 1**. That figure, and **fig. 2**, defined the physical characteristics included in the ground resistance formula.

Figure 6 shows in more detail the underground geometry and the current flow paths assumed. The ground current I_o entering the ground rod is assumed to leave the rod uniformly along its length. This is represented by the equation:

$$I = I_o \left[1 - \frac{z}{g} \right]$$

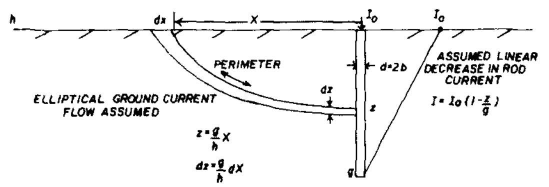

fig. 6. Ground current flow.

where z is the variable depth from the ground surface and g is the length of the ground rod.

An elliptical path is shown leaving the ground rod. An ellipse terminates at right angles to both the ground and ground rod. The path shown leaves the ground rod at z, and ends at x on the ground surface. This sheet, which starts with thickness dz, and is flattened out in **fig. 7**, carries the current dI. All such circuits are in parallel and account for the total current. For any sheet like the one shown, the current entering is:

$$dI = \frac{-I_o}{g} dz$$

As discussed earlier, the resistive current portion is:

$$dI_R = \frac{-I_o}{g\sqrt{1 + Q^2}} dz$$

The resistance to radial current flow through the parallel resistive part of the differential ring shown in **fig. 7**, is:

$$dR = \varrho \cdot \frac{length}{area} = \frac{\varrho dr}{2\pi r dz}$$

At this point, two simplifying assumptions will be made: first, the outer radius is just x rather thatn 1.1x, applicable to the elliptical perimeter, and second, that the thickness of the sheet remains a constant dz instead of increasing. These are reasonable because they introduce opposite resistance changes having a cancelling effect, and because their effect is in the far region where only a small part of the total resistance is involved. By integration, the sheet resistance to a radial current leaving the ground is:

$$dR = \frac{\varrho}{dz} \int_b^X \frac{dr}{2\pi r} = \frac{\varrho}{2\pi dz} \log_e \frac{x}{b}$$

There is an equal differential current in each sheet. Squaring the current which flows through the resistance and multiplying by the sheet resistance just arrived at, determines the sheet power:

$$dP = \left(\frac{I_o}{g\sqrt{1 + Q^2}} dz \right)^2 \frac{\varrho}{2\pi dz} \log_e \frac{x}{b}$$

Substituting

$$dz = \frac{g}{h} dx$$

$$P = \frac{I_o^2}{gh} \frac{\varrho}{2\pi(1 + Q^2)} \int_b^h \log_e \frac{x}{b} dx$$

where the quarter wave antenna height is used as a measure of the end of the outer region.

Finally, dropping the very small lower limit factor for simplicity, the approximate ground rod resistance formula is:

$$\frac{P}{I_o^2} = R_g = \frac{\varrho}{(1 + Q^2)} \frac{1}{2\pi g} \left[\left(\log_e \frac{2h}{d} \right) - 1 \right]$$

where ϱ = ground resistivity
h = quarter wave antenna height
d = ground rod diameter
g = ground rod depth
$Q = 55.6 f_{MHz} \varrho \varepsilon_r \times 10^{-6}$

appendix B
ground electromagnetic wavelength

For the Amateur HF bands, the ground is described as a quasi-conductor because of its relative dielectric constant, greater than unity.[1] In such a medium, the velocity of an electromagnetic wave is reduced, compared to the velocity of an electromagnetic wave in a vacuum. The physical characteristics which apply to a conducting medium are:

ε_r = relative permittivity (12 to 14 estimated locally)
ε_o = vacuum permittivity = 8.85×10^{-12} farads/meter
$\varepsilon = \varepsilon_o \varepsilon_r$ = actual permittivity
μ_o = vacuum permeability = $4\pi \times 10^{-7}$ henrys/meter
o = conductivity - siemens/meter
(7×10^{-3} measured locally)
$\varrho = 1/o$ = resistivity ohm-meters
$\omega = 2\pi \times$ frequency - radians/second

The equation[10] for a wave travelling in a positive x direction is:

$$E_y = E_O e^{-\gamma x}$$

where γ *is the propagation constant and* x *is the distance. The constant is:*

$$\gamma^2 = j\omega\mu o - \omega^2 \mu\varepsilon$$

It is convenient in the last term to substitute:

$$a = \varepsilon\omega/o$$

Then $\gamma^2 = \omega\mu o\,(j-a)$

But $\sqrt{j-a} = \sqrt[4]{1 + a^2}\,(\cos\theta + j\sin\theta$

where $\theta = (90° + \tan^{-1}a)/2$

Finally: $\gamma = \sqrt{\omega\mu o}\ \sqrt[4]{1 + a^2}\ (\cos\theta + j\sin\theta)$

Thus γ has a real and imaginary part. The real part is associated with attenuation and the imaginary part with phase. Equating the phase part to 2π, determines the x which is equal to one wavelength. Thus the equation for one wavelength is:

$$\lambda = \frac{2\pi}{\sqrt{\omega\mu o}} \cdot \frac{1}{\sqrt[4]{1 + a^2} \cdot \sin[(90° + \tan^{-1}a)/2]}$$

This equation has been plotted in **fig. 8**. The reciprocal of "a" is the x-axis. The label for each curve uses the variable part from the left of the fraction above. The left vertical axis is the result and gives the wavelength inside the conducting ground. It always is less than the wavelength in free space.

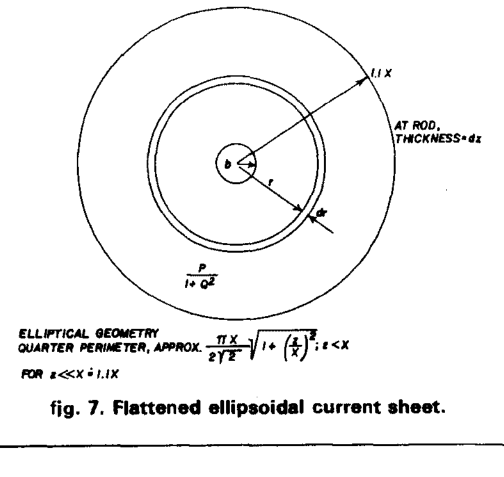

fig. 7. Flattened ellipsoidal current sheet.

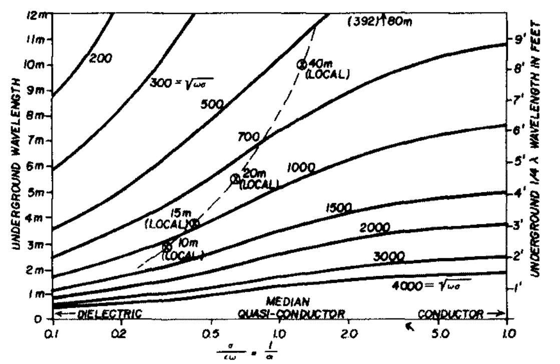

fig. 8. Quasi-conductor wavelength.

The right vertical shows the quarter wavelength in feet, which is significant for a ground rod.

On the chart in **fig. 8**, the circled crosses marked 10, 15, and so on represent the local conditions. Four-foot ground rods were used, and are very close to a quarter wavelength at 20 meters. At 10 and 15 meters they are much longer. For these, a 2-foot rod was also used.

A ground rod can be considered to be an inverted antenna in its own medium, and as an antenna, could have resonant effects. These were not included in the elementary derivation of the approximate ground resistance equation which has been presented.

Aside from any resonant effect, there is another viewpoint. The impedance of the ground rod acts as if it were inductive. The wave through the earth may be considered as made up of two parts, one through the resistive medium and the other through the dielectric medium. The latter travels at a lower velocity than the velocity through the former. The total wave is the vector sum of the two, a complex quantity which thus lags in time phase. This is analogous to the situation in a circuit having resistance and inductance in series, where the current lags the applied voltage. Therefore, the conducting medium behaves like an inductive impedance.

The existence of such an inductive impedance is supported by the frequent comment that a quarter wave antenna is much shorter than called for by the formula. A shorter antenna develops a negative reactance which neutralizes the positive ground rod reactance.

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3. *Reference Data for Radio Engineers*, Fifth Edition, 1974, ITT/Howard W. Sams, Inc., Indianapolis, Indiana, page 26-3.
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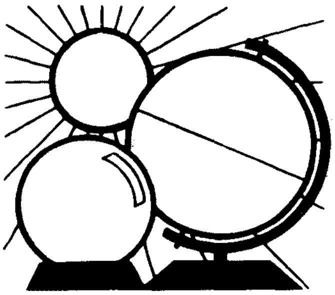
ham radio



DX FORECASTER

Garth Stonehocker, KØRYW

summer DX

Every season of the year offers advantages and disadvantages for DX operation. The winter and summer solstices (days of shortest and longest daylight) *and* the equinoxes (days of equal light and darkness) produce distinct propagation characteristics that require specialized operating procedures.

Summertime DX offers longer high frequency (10-30 meter) operating time. Higher MUFs occur earlier in the day and end later than they did in previous seasons. In the higher latitudes (i.e., in the northern hemisphere), MUFs for east-west and transpolar paths are highest at this time of year.

But summer is not without its operating disadvantages: signal strengths are generally lower, as are midday MUFs; hop lengths are shorter; and there's little one-long-hop transequatorial propagation. Amateurs in the northern hemisphere can expect increased noise QRN as well.

Some of these negative characteristics of summertime operation can be turned into advantages. Others can be overcome. In this column and the next, I'll discuss these characteristics and offer some suggestions for improving your summertime DX operating.

last-minute forecast

The higher frequency bands (6-30 meters) are expected to be best during the first ten days of July. During this time, both long and short skip openings should be plentiful except on 6 meters, where only short skip openings by sporadic E are possible. During the latter two weeks of the month sporadic-E short skip will provide the best openings. The lower bands (30-160 meters) are expected to be of limited use during the month because of summer thunderstorm QRN; the first of the month may have the better nighttime openings, and the last two weeks the better daytime openings if the solar radio flux is 80 or less, as expected.

A full moon will occur on the 13th; perigee (closest approach) is on the 2nd and 30th. The Aquarid meteor shower starts on July 18, peaks on the 28th, and lasts until August 7 (all dates approximate, but close). The radio-echo rate at maximum is about 34 per hour.

band-by-band summary

Six meter paths will exist during several days around local noon by means of sporadic-E short skip propagation.

Ten and fifteen meters will provide long-skip conditions in the afternoon during the peak times of the 27-day solar maximum. Otherwise, look to sporadic-E short-skip and multihop openings around local noon for DX on these bands. Transequatorial evening openings do not usually occur in the summertime.

Twenty and thirty meters will be open all day and much of the night. If twenty does not stay open through the night, thirty probably will. Sporadic-E short-skip is also often effective on these bands throughout the day. Propagation paths to most areas of the world are possible in a sequence that follows the sun's journey across the sky: east in the morning, south during midday, and west during the evening.

Thirty and forty meters will be the main nighttime DX bands this time of year, although long-skip distances will be shorter. Sporadic-E openings are possible during most of the day into pre-sunrise and after sunset. With thunderstorm-induced static levels high in the evening, look to pre-dawn periods for best results.

Eighty and one-sixty meters are difficult DX bands during this time of year. Short nights and high noise levels hamper DX operation, although eighty offers slightly lower noise levels. Most useful openings may occur during the pre-dawn hours. Sporadic-E propagation signal strengths may exceed the static level near sunrise and sunset.

ham radio

WESTERN USA

GMT	PDT	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)
0000	5:00	20	20	15	10	20	10	10	20
0100	8:00	20	20	20*	10	20	10	10	20*
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0300	8:00	20	20	20	10	20	10	10	15
0400	9:00	20	20	20	10	20	15	10	15
0500	10:00	20	20	20	10	20	15	15	15
0600	11:00	20	_30*_	20	15	30	15	15	15
0700	12:00	20*	_30*_	_20_	15	30	15	15	15
0800	1:00	20*	_30*_	_20_	_15_	_30_	20	15	20
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1700	10:00	20	20	10	20	20	15	20	20
1800	11:00	20	20	10	20	20	15	20*	20
1900	12:00	20	20	10	15	20	15	15	20
2000	1:00	20	20	10	15	20	15*	15	20
2100	2:00	20	20	10	15	20	10	15	20
2200	3:00	20	20	15	15	20	10	15	20
2300	4:00	20	20	15	15	20	10	15	20

JULY

MID USA

GMT	MDT	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)	CDT
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0600	12:00	20	_30*_	20	20	30	15	15	20	1:00
0700	1:00	_20_	_30*_	_20_	20	_30_	20	15	20	2:00
0800	2:00	_20_	_30*_	_20_	_20_	_30_	20	15	_20_	3:00
0900	3:00	_20_	_30*_	_20_	_20_	_30_	_20_	20*	_20_	4:00
1000	4:00	_20_	_30*_	_20_	_20_	_30_	_20_	_20_	_20_	5:00
1100	5:00	_20_	_20_	_15_	_15_	_30_	_20_	_20_	_20_	6:00
1200	6:00	_20_	20	15	15	_30_	_20_	_20_	_20_	7:00
1300	7:00	20	20	15	15	_30_	_20_	_20_	_20_	8:00
1400	8:00	20	20	15	15	_30_	_20_	_20_	_20_	9:00
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1700	11:00	20	20	10	10	20	15	20	20	12:00
1800	12:00	20	20	10	10	20	15	20	20	1:00
1900	1:00	20	20	10	10	20	15	15	20	2:00
2000	2:00	20	20	10	10	20	15	15	20	3:00
2100	3:00	20	20	10	10	20	10	15	20	4:00
2200	4:00	20	20	15	15	20	10	15	20	5:00
2300	5:00	20	20	15	15	20	10	15	20	6:00

EASTERN USA

GMT	EDT	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (Caribbean S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)
0000	8:00	20	20	15	15	20	10	10	20
0100	9:00	20	20	20*	20	20	10	10	20
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0500	1:00	20	20	20	20	30	15	15	20*
0600	2:00	_20_	_20_	20	20	30	15	15	20
0700	3:00	_20_	_30*_	20	20	30	20*	15	20
0800	4:00	_20_	_30*_	_20_	_20_	30	20	20*	_20_
0900	5:00	_20_	_30*_	_20_	_20_	_30_	20	_20_	_20_
1000	6:00	_20_	_30*_	_20_	_20_	_30_	_20_	_20_	_20_
1100	7:00	_20_	_30*_	_15_	_15_	_30_	_20_	_20_	_20_
1200	8:00	_20_	20	15	_15_	_30_	_20_	_20_	_20_
1300	9:00	20	20	15	15	_30_	_20_	_20_	_20_
1400	10:00	20	20	15	15	_30_	_20_	_20_	_20_
1500	11:00	20	20	15	15	_30_	_20_	20	_20_
1800	12:00	20	20	10	10	20	15	20	20
1700	1:00	20	20	10	10	20	15	20	20
1800	2:00	20	20	10	10	20	15	20	20
1900	3:00	20	20	10	10	20	15	15	20
2000	4:00	20	20	10	10	20	15*	15	20
2100	5:00	20	20	10	10	20	10	15	20
2200	6:00	20	20	15	15	20	10	15	20
2300	7:00	20	20	15	15	20	10	15	20

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during "normal" hours.
*Look at next **higher** band for possible openings.

the ham notebook

RIT for the KWM2

Tired of playing leapfrog when working stations on CW that have rigs like yours, but lack any clarifier provision in the receive mode? And are you tired of working in nets where 20 percent of the operators aren't exactly on frequency, and you have to keep leaving the NCS (Net Control Station) frequency in order to read them clearly? Well, I was — so I decided to add a clarifier module to my KWM2A (**fig. 1**).

Most of the necessary parts, with the exception of a varactor diode, are probably already in your junk box; the diodes are quite plentiful and reasonably priced.

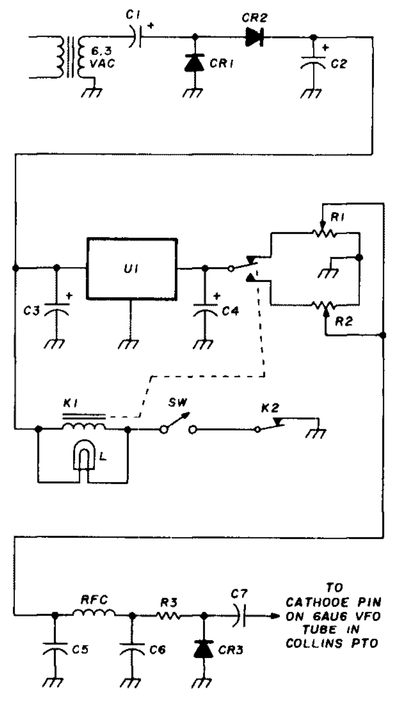
design approach

No kinks were found in this unit; it works like a charm, providing about 2 kHz or better frequency swing on receive.

You can make the swing as much as you desire by using different types of varactor diodes or by varying the size of C7 (see **fig. 1**), which is effectively in series with the varactor diode. The diode I used has a capacitance variation of 4.5 to 22 pF when the bias is varied from 0 to 10 volts. Many varactor sizes are available from 5 to 10 pF up to hundreds of pF in value. Choose whatever you need; any of those in the parts list should suffice.

For the Collins modification I decided the best place for the power supply section was in the external power supply speaker unit. I had never used the built-in speaker, so I just removed it and used the space to house a board with the voltage doubler circuit shown, using the 6.3 VAC filament supply as a source.

I found one lead in the power cable from the AC power supply to the KWM2 that terminated in an unused pin (power supply end). This is pin 6, which actually has switched 110 VAC coming from the transceiver end. I disconnected the 110 VAC from this lead and used it to carry the 15 VDC from the AC power supply into the transceiver. I also disconnected this lead from the T/R relay, K4, in the KWM2, in which it formerly appeared on relay contact No. 13, and ran it via shielded wire up through a hole in the chassis near the VFO unit. This becomes the 15 VDC supply to power the RIT board and regulator circuit. This leaves a relay arm and two contacts (No. 12 and 11) unused in T/R relay K4. I put the arm (No. 12) to use as a disable circuit so that during transmit the clarifier is automatically out of the circuit, regardless of whether it is left switched on from the front panel switch or not. When receiving, this relay arm furnishes a ground and in turn activates the RIT control relay. When transmitting, however, this arm closes to contact No. 13; this prevents the confusion that would result from transmitting with the clarifier inadvertently left on and thereby operating on an undesired frequency. I made the entire regulator and RIT circuitry on one small perforated hole board and mounted it on the two screws on the back of the Collins PTO. This puts it as close to the VFO tube (6AU6) as possible and permits running a short lead from the varactor diode blocking capacitor, C7, to the cathode lead of the 6AU6 VFO tube. It isn't necessary to enter the PTO box at all. Just wrap a small solid wire a turn or two around the cathode pin of the 6AU6 tube and solder it. Be careful; too much heat for too long a time will cause the glass on the tube to crack. Be sure the potentiometers used are the linear taper type in order to obtain a smooth frequency swing from stop to stop; one is used for adjusting the transmit frequency and should be set somewhere near mid-range. The other potentiometer is switched in only when the RIT switch is on. This potentiometer has a shaft brought out to the front panel right over the top of the PTO chassis. This is your receiving RIT control. If you're squeamish about drilling holes in the front panel, knocking out the "Collins" name decal will leave a hole in the



item	description
C1,C2	500 microfarad 25 volts
C3,C4	100 microfarad, 25 volts
C5,C6	0.01 microfarad disc ceramic
C7	40 to 80 picofarad NPO
CR1,CR2	1 amp 50 volt diodes
CR3	varicap diode 1S2236
K1	SPDT small 12 volt relay (low current type)
K2	contacts on relay K4 inside KWM2
LAMP	small 12 volt low current pilot lamp
R1,R2	2.5K linear potentiometer
R3	10K 1/2 watt
RFC	any small RF choke or ferrite bead
SW	SPST mini toggle switch
U1	3-terminal 12 volt regulator

Notes: the value of C7 will depend on how much frequency swing one desires for the RIT function. The smaller amount gives about 1.5 kHz spread. The larger amount up to 3 kHz spread. I used a 1S2236 (Japanese part number), but any varicap that will give a swing from about 10 to 30 pF will work. Others that would be satisfactory are the 1N950 and 1N955.

panel through which the control shaft may be passed.

I put the on/off switch and an optional warning light (to let me know when the clarifier is operating while in receive mode) below the PTO dial area near the bottom of the front panel near the center. You can use the new mini-toggles or a push-on/push-off mini-switch and one of the smaller types of pilot lights. Make sure the relay, K1, and the pilot lamp are both low-current consumption devices. If these two items pull too much current from the unregulated 15 VDC source, degradation of the voltage regulator's operation could result.

My relay is very small and has the required SPST configuration. The relay and the pilot lamp pull only about 50 mA of current. I would say that up to 100 mA could be handled without any loading down of the 15 VDC supply. The operation of the clarifier is straightforward, and has no adverse effect on the stability of the VFO; whatever stability existed before will remain even with the clarifier in use.

I happened to use a 12-volt three-terminal regulator because that was what I found in my junk box, but you could use a 9-volt unit as well, except for its slightly narrower frequency range on the RIT. Also, when reconfiguring the power lead pins, I found that C108 and R101, which were formerly across the K4 relay contacts, were no longer needed, so I clipped them out. (Tracing these two components in the KWM2 is easy as they are right behind the K4 contacts in question. Note also that the K4 contact numbering is different in the older KWM2 *than in the later KWM2A series.* On the former, the contacts were numbered 9, 10, and 11, instead of 11, 12, and 13.)

not limited to Collins transceivers

This modification can be used in most transceivers that do not have RIT, such as the Drake TR4, Heath SB104, etc., by connecting a lead from the varactor blocking capacitor C7 to the proper point in your VFO tuned circuit. This additional capacitance helps tune the VFO. This in turn will require you to estimate the amount of capacitance your varactor diode will need in order to cause your VFO frequency to swing the desired number of kHz.

After adding this modification to the KWM2, your VFO (Collins PTO) will read a little off frequency one way or the other. Just take a turn or two on the coil slug, which is on top of the PTO box and just to the right of the 6AU6 VFO tube. This adjustment corrects the readout for all bands. The amount you have to adjust this slug is small and will depend somewhat on the particular varactor diode you select.

Joe Fenn, KH6JF

QRP wattmeter

Prowling around radio swap meets or surplus stores, one sometimes finds a Weston Thermo Galvanometer. Because no one seems to know how they can be used, they just stay in the junk pile.

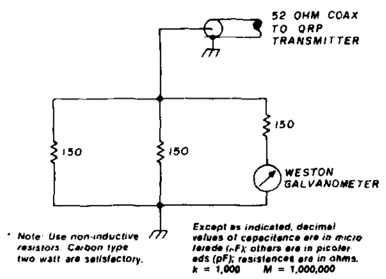


fig. 1. QRP wattmeter schematic.

A 0-100 mA RF current galvanometer makes an excellent direct-reading 0-5 watt power meter that can be used with QRP rigs. The internal resistance is 5.2 ohms for 100 mA full scale reading.

Using three 150-ohm 2-watt resistors in parallel, and putting the galvanometer in the leg of one of them makes an excellent wattmeter (**fig. 1**) with an accuracy of 2 percent up to 65 MHz.

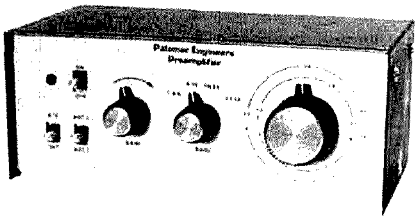
Ed Marriner, W6XM

book and product

REVIEWS

Palomar Engineers P-308 preamplifier

Not everyone needs a preamplifier these days because most of the newest generation of receivers and transceivers have more than sufficient sensitivity. Some, like the ICOM 751, Yaesu FT102 and TenTec Corsair, to name a few, have switchable RF amplifiers already installed. However, for older solid-state units and for tube-type radios, preamplifiers can make the difference between being able to hear the weak station or missing it altogether.



Another situation that will often require the use of preamplifiers is when you're using specially designed receiving antennas such as beverages and loops. These antennas aren't "gain" devices and are, in fact, 10-20 dB down from full-sized dipoles or other types of antennas. They're used because they're less susceptible to noise and because they exhibit directivity.

In my installation, the beverages are several hundred feet away from the station, so in addition to the low signal input, I also have several hundred feet of coax to get a signal through. As you VHF/UHFers know, this is a major problem. On 160 meters, coax loss isn't nearly as bad as at VHF/UHF, but a dB here, a dB there, all add up.

I looked through my 160-meter file to see what I had in the way of information on preamps and found several interesting circuits and specs for commercially manufactured gear. I realized I wouldn't have enough time to collect all the parts I needed to complete the project before the upcoming 160 meter CQ CW test. So I decided that this would be a good opportunity to try out a commercially manufactured preamp, and I gave Palomar Engineers a call.

Within a few days I had a P-308 preamplifier in hand to try out.

the unit

The P-308 is one of four dual-gate FET amplifiers available from Palomar. The P-308 is designed to be used with receivers and runs off 110 VAC (the P-305 is the same unit, but it runs off a 9-volt battery). The P-310X and P-312X are designed to be used with transceivers and have a built-in carrier operated relay (COR) that automatically switch the preamp circuit in and out.

All Palomar preamps use a tuned RF circuit and cover 1.8-54 MHz in four continuous bands: 1.8-4 MHz, 4-10 MHz, 10-23 MHz, and 23-54 MHz. The preamp uses a dual-gate FET and has a variable gain of up to 20 dB controlled by a front-panel control. The unit also has a 20-dB pad included for situations in which you might have too much gain in crowded band conditions, where cross mod or overloading the receiver could be a problem.

One tricky feature about the unit is making absolutely sure that it is properly tuned when in use. Failure to properly tune the preamp can result in images or spurs being amplified at the same time as the weak signal. The RF tuner stage uses a tapped coil that is switch-selectable by band and a 150-pF variable capacitor. Careful adjustment of the capacitor will eliminate this potential problem.

use

I've used this unit during the last two 160-meter contests and found it to be quite helpful in bringing those very weak DX signals up to an audible level. (That doesn't mean I actually *worked* them — but at least I could hear them!) During the contests, careful adjustment of the capacitor and gain reduction reduced or eliminated cross mod and overloading, due to the extremely crowded band conditions. (Last night I tried to get through the "wall of 4s" to work YV0AA. The YV0 was uncopyable without the preamp being on.)

I haven't spent much time with the P-308 on 15 or 10 meters. I would expect that it would be useful at that end of the ham spectrum, because that's the region where many ham receivers and transceivers need extra "oomph."

conclusion

If you're looking to pep up an old tube-type receiver, one of the Palomar preamplifiers may be just right for you. The attractive silver-gray front panel and black chassis will fit quite nicely into any ham shack. The unit weighs approximately 3 pounds and measures 8 × 5 × 3 inches.

For the name of your local dealer, contact Palomar Engineers, Box 455, Escondido, California 92025.

Circle #315 on Reader Service Card.

N1ACH

"The Complete DX'er" by Bob Locher, W9KNI

Over the years there have been some pretty good DX books, and a few that have been simply terrible. What there hasn't been is a really great DX book — until now, that is. Drawing on the skill and knowledge that earned him one of the first CW DXCCs in 1975 and has kept him at the top of that mode's DXCC Honor Roll ever since, W9KNI has created a first-class treatise on the art of DXing that's every bit as entertaining as it is educational.

Of course this should be no surpise to readers of *ham radio* or *Ham Radio Horizons*. Bob's very popular articles on DXing and DX adventures in the two magazines were avidly read by all kinds of people, even non-hams, because they were well written and entertaining. Hams read them eagerly because each one of those articles included priceless "tricks of the trade" known but to consumate DXers. No one, whether newly-licensed Novice or grizzled DXCC Honor Roll member will fail to find something useful in Bob's latest dissertation on DXing. (Even though Bob's credentials are for CW, what he has to say applies equally well to other modes as well. That's why this is a unique DX book.)

The book is divided into two sections, the first primarily directed to the beginning DXer and the second for the advanced (200+ countries) DX chaser. The experienced DXer might be tempted to skim or even skip the first. *Don't*. It's not only very entertaining reading, but every page is laced with those little morsels of DX wisdom that always spice Bob's writing. Oh yes, there are chapters on the basics of listening, operating, and equipment, but even in those mundane subjects Bob has sewn some pearls that almost any DXer, regardless of experience, will find profitable.

The fun continues in the second half. After a very practical discussion of the art of QSL-chasing, Bob launches into "Graduate Hunting" and later on "Hunting — Again." Here the reader learns that a first class station and operating skills are not always sufficient in themselves. The top DXer must also bring to the chase the deductive ability of a Sherlock Holmes, the results of which Bob demonstrates with some fascinating anecdotes. Finally, after a much-needed chapter on the ethics of DXing and discussions on various special tools for the DXer, Bob sends the reader back to the shack to resume the search for that ever-elusive new one. He'll find him, too, now that he's been helped along by one of the true masters of the art.

Those who buy this book looking for beam heading charts, prefix lists and postal rates are going to be disappointed — there are no charts or tables. What it does contain is the wit and wisdom of DX chasing, written by an acknowledged expert. This book could be the best investment a DXer could make; I can think of no higher recommendation!

The Complete DX'er is available from Ham Radio's Bookstore, Greenville, NH 03048, for $10.95 plus $2.50 shipping and handling.

W9JUV

NEW products

ICOM mobile transceiver

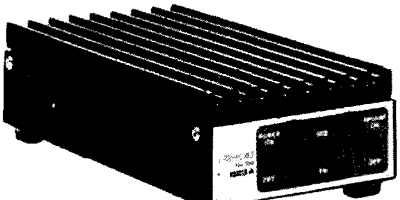
ICOM has announced the release of the new IC-37A 220 MHz ultra-compact mobile transceiver. The IC-37A features: 25 watts/5 watts low; and measures only 5-1/2″ wide × 1-1/2″ high × 7″ deep, offering the same compact design as the IC-27A. 32 PL frequencies are standard and built-in; nine memories with offset and PL storage are included. Dial steps are in increments of 10 kHz/5 kHz. Scanning features include memory scan, band scan, and priority scan. Dual VFOs and HM-23 Touchtone® and scanning mic are standard. A speech synthesizer option is available.

For additional information, contact ICOM America, Inc., 2112 116th Avenue, N.E., Bellevue, Washington 98004.

Circle #317 on Reader Service Card.

H/T amplifiers

Mirage Communications has added two new low-profile H/T amplifiers to its expanding line of American made communications equipment.



The B23A (144-148 MHz) and C22A (220-225 MHz) incorporate features that are typically available only on larger, more expensive amplifiers. Both feature a built-in receive preamp that delivers a 1.5-2.0 dB noise figure, all-mode operation (CW, FM, or SSB), and automatic antenna changeover.

Their RF power input range is from 100 mW to 5 watts; high RF power output (B23A, 2W in, 20W out/C22A, 2W in, 30W out) makes them ideal for use with virtually all low-power transmitters.

They're each backed by a five-year factory warranty (1 year on RF power transistors) and a worldwide sales network, and priced at $89.95.

For more information, contact Mirage Communications, P.O. Box 1000, Morgan Hill, California 95037.

Circle #313 on Reader Service Card.

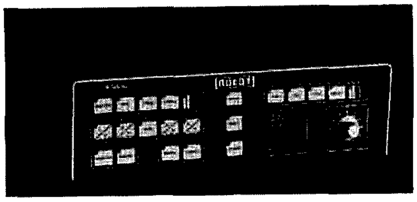
NEW products

robot color SSTV converter

Robot Research has introduced a new black and white-compatible single-frame slow scan TV color converter. Designated Model No. 450C, it incorporates the latest in microcircuit technology and represents a significant breakthrough in the application of microprocessor-based systems to Amateur SSTV.

Model 450C features touch-sensitive front panel switches for full station control and several automatic functions for easy use. Fine tuning, speed switching and color or black and white detection are automatically accomplished without operator intervention.

Three selectable 4-bit memory planes combine to form 4,096 color combinations in a 128 pixel by 120 line full screen display. Eight different black and white and color transmission formats are available with automatic selection on receive. The unit accepts color or black and white composite video from standard TV cameras and has RGB, composite or RF modulated video output.



A unique feature of the 450C is the 8-bit parallel I/O ports for computer interface and hard copy picture printing. This analog provides total access to each individual pixel by a host computer for image processing, transformation, storage and recall, and graphics. In addition, the unit has provisions to connect to the new robot model 800C super terminal for color graphics, SSTV graphics composition, graphics overlays, and special test patterns.

For further information, contact Robot Research, Inc. 7591 Convoy Court, San Diego, California 92111.

Circle #312 on Reader Service Card.

Model 230 decoder

SYT Corporation has announced the Model 230 DTMF mobile decoder for autopatch and selective call systems.

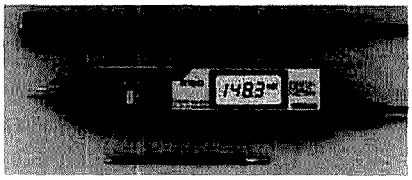
The 230 DTMF decoder is housed in a compact metal housing for RFI protection. It features a high-speed precision decoder with wrong-digit lockout to prevent falsing. It is field programmable for DTMF code lengths from 1 to 8 digits. Call alert outputs are provided for horn, external speaker control, and internal call lamp. An optional internal alert buzzer is available.

For more information, contact Larry Francis at SYT Corporation, 1220 Barranca, El Paso, Texas 79935.

Circle #311 on Reader Service Card.

pen-type DMM

North American Soar Corporation has released the new Model 3100 Pen Type DMM, offering 25 percent greater range capability and a 3 1/2 digit LCD readout with annunciators that are said to be 50 percent larger than similar units presently available.



Model 3100 is fully autoranging and can measure DC volts from 0.1 mV through 500 volts on five ranges, AC volts from 1 mV through 500 volts on four ranges and resistance from 0.1 ohm through 20 Megohms on six auto-ranges. This dynamic range capability is made possible by use of Soar's custom 80-pin flat pack chip. Use of this state-of-the-art circuit allows fast, stable readings to be obtained. The Model 3100 has an audible continuity check feature that indicates a circuit is continuously conductive by beeping; the beeper reaction time is 0.4 seconds or less. A "Data Hold" switch is provided for freezing a reading even though the test probe is removed from the test point.

The unit is supplied with two interchangable test points, one uninsulated 1/2 inch and one 2 3/4 inch insulated extension tip, safety type ground probe with slip-on insulated alligator clip, and two batteries. The Model 3100 is priced at $49.99.

For further information, contact North American Soar Corporation, 1126 Cornell Avenue, Cherry Hill, New Jersey 08002.

Circle #310 on Reader Service Card.

Transi-Trap™

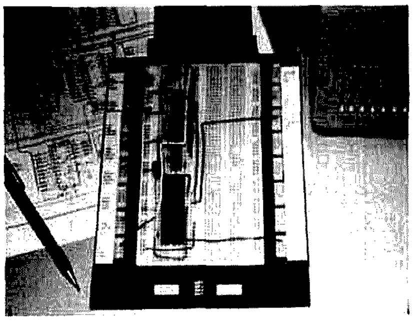
The new Alpha Delta Model ACTT AC Transi-Trap™ is a direct plug-in-the-wall AC surge protector providing two 120 VAC sockets, status light, circuit breaker, and a three-stage automatic surge protection circuit. The Model ACTT provides both transverse and common mode protection with a hot-to-neutral,, neutral-to-ground and hot-to-ground 6000 volt, 2000 ampere surge discharge self-restoring high speed circuit. The configuration of the Model ACTT also protects equipment plugged into any other common branch AC wall outlet down line from the ACTT. The unit is UL listed and is priced at $29.95.

For further information, contact Alpha Delta Communications, P.O. Box 571, Centerville, Ohio 45459.

Circle #307 on Reader Service Card.

computer prototyping board

The "eZ BOARD" is a solderless system for building experimental add-ons to interface with personal computers. Its features include a glass epoxy printed-circuit board mounted with a set of solderless breadboarding units. Four separate distribution buses with fifty tie-points each can be used for power, ground, clock lines, reset commands, and other purposes. A four-position DIP switch is mounted on the board. Each switch position connects to a set of the tie-point block sockets on either side. A flat ribbon cable connects the board to the computer's bus expansion slot. The breadboarding area consists of 1460 tie-points with a capacity of sixteen 14-pin DIPs. Components with lead diameters up to 0.032 can be plugged in with ordinary solid hookup wire.



Models IPC, APC, and CPC are available for IBM-PC, APPLE and Commodore, respectively, and all other hardware-compatible computers of similar types. Models for other computers will be introduced during 1984. The retail price for the entire system, including cable and connectors, is $174.95 plus $5.00 shipping.

For further information, contact Sabadia Export Corporation, P.O. Box 1132, Yorba Linda, California 92686.

Circle #308 on Reader Service Card.

VHF/UHF diplexer

If you operate VHF and UHF, you know that long coaxial cable runs will reduce your power to the antenna significantly. You also know that low-loss coaxial cable is quite expensive. So what do you do if you want to operate both VHF and UHF without using a lot of expensive coax to feed your antennas? Microwave Filter has come up with an answer: the Model 4525 diplexer.

But what's a diplexer? A diplexer either combines or splits RF signals; for example, you could use one diplexer in the shack to combine RF from your 144 MHz and your 420 MHz radios. These

combined signals would then be fed through just one coaxial cable to the top of the tower, where a second diplexer would be located. This diplexer would again split the signals and route one to the 144 MHz antenna and the other to the 420 MHz antenna.

The Model 4525 diplexer, priced at $104.50, is designed for the 144 and 420 MHz bands. With insertion loss less than 0.4 dB, it's designed to handle up to 200 watts of RF.

For more information, contact Microwave Filter Company, 6743 Kinne Street, East Syracuse, New York 13057.

Circle #309 on Reader Service Card.

Design Kit™ Series

Communications Consulting Corp. has announced the availability of its General Purpose Design Kit™ series. The series currently consists of an RF design kit, communications design kit, and a PLL design kit.

Most software packages available for desktop computers are concentrated in the area of computer aided analysis. The important area of circuit *synthesis* is frequently overlooked. In introducing the Design Kit™ Series, Communications Consulting Corp. has made these very powerful tools available for the HP series 9845B/C, the HP 9000, series 200 and series 500 (Model 9020). They all consist of a universal mathematical function plot and a frequency selection for minimum IMD products in mixers and relevant programs.

The RF Design Kit™ includes software for a universal mathematical function plot; frequency selection for minimum IMD products in mixers; optimization of noise figure or intercept point of cascaded amplifiers, mixers, and filters; determination of noise figure from S/N ratio; compensated wideband transformer; equalizers; quarter and half wavelength UHF oscillators; and design of microwave striplines.

The Communications Design Kit™ contains software for a universal mathematical function plot; frequency selection for minimum IMD products in mixers; dual loop AGC with group delay correction; digital filter design program; complex impedance of electrical short antennas; antenna array pattern with all driven elements; and quarter and half wavelength UHF oscillators.

The PLL Design Kit™ includes software for a universal mathematical function plot; frequency selection for minimum IMD products in mixers; optimization of type 2, second order loops; optimization of type 2, third order loops; optimization of type 2, fifth order loops (these programs include BODE plot, lock-up calculations, and component selection); general analysis of PLL circuits; and quarter and half wavelength UHF oscillators.

For further information, contact Communications Consulting Corp., 52 Hillcrest Drive, Upper Saddle River, New Jersey 07458.

Circle #306 on Reader Service Card.

Design Kit™ is a registered trademark of Communications Consulting Corp.

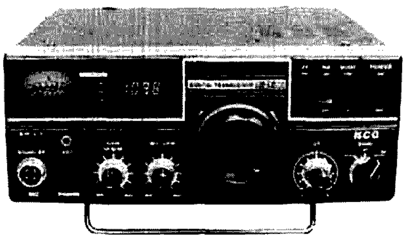
✓ 177

NEW products

7-21-50 radio

The new 7-21-50 from NCG is an all solid-state radio that covers the complete 40 and 15-meter bands and from 50 to 50.5 MHz on the 6-meter band. All bands operate in either the SSB or CW mode; PLL circuitry in the VFO ensures drift-free operation. The transmitter runs 26 watts PEP and utilizes balanced modulation for clear, crisp SSB audio.

Designed with extensive TVI suppression to minimize TVI problems while operating on a well matched 6-meter antenna, the 7-21-50 is small enough to be used mobile or portable, and offers additional features for use as a base station radio,



too. No external power supply is required. Technician class hams will find the 7-21-50 to be well suited to 6 meter SSB and 40 and 15 meter CW.

The 7-21-50 comes complete with an internal AC/DC power supply.

For additional information, contact NCG, 1275 N. Grove Steet, Anaheim, California 92806.

Circle #316 on Reader Service Card.

computer-controlled scanner

Electra Company has announced that the Bearcat® CP-2100, said to be the first scanner radio designed as a peripheral for today's popular personal computers, is now available in versions compatible with the IBM Personal Computer, Atari 800, Apple II and IIe, Osborne and Commodore 64 personal computers.

The CP-2100 can monitor live police and fire calls, emergency and Amateur Radio transmissions, Coast Guard rescues, and aircraft communications.

Each of its 200 channels can be programmed to display the source and location of a transmission, 10-codes, phone numbers and more. Whenever a broadcast is monitored, the information progammed into the channel will automatically appear on the screen. For scanner enthusiasts, the Bearcat CP-2100 can eliminate pages of cumbersome frequency lists. In the newsroom, it can help news crews be dispatched to the scene of a story with less confusion.

Its 200-channel capacity is three times that of the most sophisticated programmable scanners.

The Bearcat CP-2100 is also the first scanner to feature multiple priority levels. Users can select up to three different priority frequencies so most important calls will be heard first. During a priority transmission, the video monitor will flash to alert the listener. "Search/Store/Count" lets users search frequency ranges of their choice for active channels. The scanner will automatically find all active frequencies and store them in separate memory without the need for the operator to be present. The count register shows how many transmissions were noted on each active frequency. In addition, every channel includes four auxiliary settings which can be programmed to activate tape recorders, alarms and other optional equipment whenever a call is received on the programmed channel.

Other features include patented "selective scan delay," "automatic lockout," "automatic and manual search," and patented "track tuning." Frequency coverage includes 10, 6, and 2-meters, and 70-centimeter Amateur; VHF Low and High; VHF Aircraft; UHF and UHF-"T"; and Military Land Mobile bands.

The Bearcat CP-2100's basic package includes the radio, AC adapter, plus a special telescoping whip antenna with 20 foot coaxial cable and mating BNC connectors. A custom 5-1/4 inch program diskette, custom interface cable, and manual are packaged separately for compatability with different models of computers. The software can be user-modified to suit individual needs.

The suggested retail price of the Bearcat CP-2100, including both hardware and software packages, is $499.95. For additional information, write Electra Company, 300 East County Line Road, Cumberland, Indiana 46229.

Circle #304 on Reader Service Card.

VIC-20 programs

Three new HAMWARE programs developed by John Vesty Company are said to extend the utility of VIC-20 computers to logging and QSO operations.

Ham List serves as a memory jogger during a QSO, quickly searching for a call and displaying data on file. The program provides for the convenient addition, revision, or deletion of entries, and a screen-review of the list.

Quick Log provides automatic logging of data and time, and search by call or QTH. The list can be printed, saved to tape, or screen-reviewed as desired. Time is displayed on the menu page.

QSO Manager combines a ten-minute identification timer and a 24-hour clock, with a screen-based notepad for use during a phone or CW QSO. The notepad incorporates a word-wrap routine to eliminate broken words at the end of a line. The timer can be set, reset, or cancelled at any time.

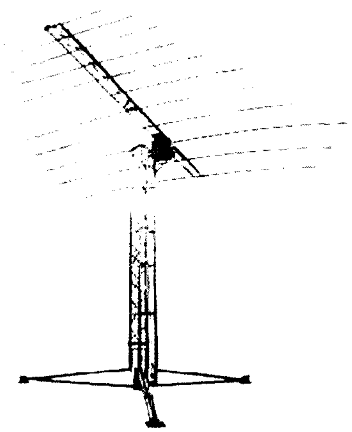
The three programs are available on tapes, and are designed for use with both unexpanded and expanded VIC-20 computers. The capacity of the logging programs ranges from 100 to 700 entries maximum, depending on the memory expansion used and the length of individual entries.

For further information, contact John Vesty Company, 415 Elm Street, Fayetteville, New York 13066.

Circle #303 on Reader Service Card.

support system

"Lightfoot," a new antenna support system that evenly distributes loads to rooftop structures, features a pedestal that holds a tower/elevator suspended above the roof with three extended legs secured to the roof structure on mounting pads at the outer ends.



This design is said to reduce roof load by as much as 65 percent over conventional rooftop systems.

Once in place, an elevator track allows the array or rotator to be lowered to comfortable working height by one person in minutes. The height is 23 feet, pedestal radius is 13 feet. The aluminum unit weight is 640 pounds; the steel unit, 1400 pounds.

Designated RM-20, the system is available in either steel or aluminum. Additionally, the RM-20 is available in combination with antenna arrays and rotators, as well as a power winch system for raising and lowering.

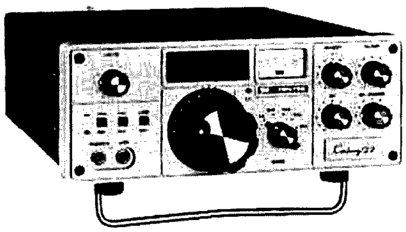
For further information, contact Sabre Communications Corp., 117 Main Street, Sioux City, Iowa 51102.

Circle #302 on Reader Service Card.

NEW products

Ten-Tec Century 22

Ten-Tec has just released its newest radio, the Century 22. The Century 22, an update of the popular Century 21, features the latest in solid-state RF technology. The design concept is a low power, gimmick-free, CW only radio. Ideal for beginners on a budget — as well as for established hams as either a second set or mobile radio — this new unit offers full QSK, 50 watts of RF input power, a six-pole audio lilter, linear crystal mixed VFO and covers the standard five bands 80-10 meters plus the WARC 30-meter band.



The Century 22 also features receiver sensitivity of 0.5 μV, offset receiver tuning, automatic gain and level control, adjustable side tone, CW and SSB or receive, and SWR and forward power metering. The unit weighs just 6 pounds (2.7 kg) and measures 4 x 10 x 10.5 inches (10.2 x 25.4 x 26.67 cm). It has a rigid metal chassis with molded front panel and aluminum back, top and bottom. It includes 28 transistors, 24 diodes, and 7 ICs. Frequency tolerance is $\pm$ 5 KC of dial reading, ($\pm$2 KC when using the optional calibrator). The Century 22 is priced to fit the beginner's budget at $389. The optional 110 VAC to 12 VDC, 5-amp power supply is $89 additional.

For more information, contact Ten-Tec, Inc. Sevierville, Tennessee 37862.

increased weather protection

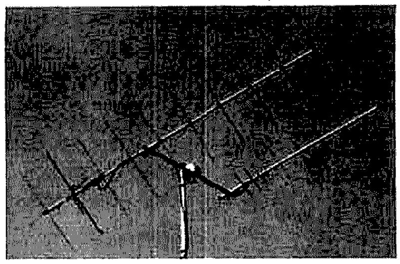
The improved NMO series of antenna hardware from Larsen Electronics features an extra lip around the bottom of the coil, providing a place lor an "O" ring. The ring surrounds the threads that tighten to the vehicle surface, eliminating the chance that water may be drawn through the threads, decreasing performance and corroding mounting hardware. This improvement is said to make a difference in areas where antennas are constantly exposed to heavy rain, and to offer additional protection under less severe conditions as well. All NMO-series VHF and UHF products now incorporate this change; LM and NLA series are also equipped with a neoprene gasket around mounting hardware for a weather-tight seal.

For details, contact Larsen Electronics, P.O. Box 1799, Vancouver, Washington 98668.

Circle #314 on Reader Service Card.

satellite antenna

Cushcraft has introduced a new Amateur satellite antenna system featuring two high gain circularly polarized Yagi antennas. The 70 cm, 16-element uplink and 2-meter, 20-element downlink antennas are fixed to a common mounting boom. The entire array is lightweight with reasonable dimensions for quick installation.



For more information, contact Cushcraft Corporation, P.O. Box 4680, Manchester, New Hampshire 03108.

Circle #301 on Reader Service Card.

HF antenna design program

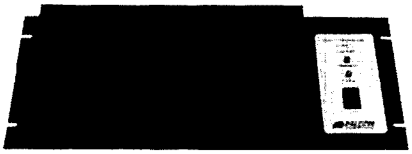
HF Antenna Design is the latest offering in Cynwyn's line of software for Amateur Radio operators. The program makes the necessary calculations for building three popular types of antennas — dipole, Yagi, and quad — for frequencies of 1.8 - 30 MHz and displays them in an easy-to-read tabular format. The dimensions for the Yagi and quad are optimized for maximum gain.

HF Antenna Design required a TRS-80C Color Computer^{TM} with 16K RAM and Extended Color Basic or MC-10 with 4K RAM. The program is available from Cynwyn — 4791 Broadway, Suite 2F, New York, New York 10034 — on cassette for $10 plus $2 shipping and handling.

Circle #305 on Reader Service Card.

repeater amplifier

Falcon Communications has announced its new Model 4114 2-meter repeater amplifier, a basic amplifier that supplies a full 100-watt output when driven with 2 watts. Features include carrier operated relay or external keying; operation on 13.8 volts DC, from either a battery or power supply; regulated bias supply (adjustable for limited power output adjustment); and a thermostat designed to prevent any damage in the event of overheating. No damage results from high VSWR.



Other units in this series supply maximum power outputs of from 50 to 100 watts and accept drive levels up to 25 watts. A 220 MHz unit is also available.

For further information, contact Falcon Communications, P.O. Box 620625, Woodside, California 94062.

flea market

RATES Noncommercial ads 10¢ per word; commercial ads 60¢ per word **both payable in advance**. No cash discounts or agency commissions allowed.

HAMFESTS Sponsored by non-profit organizations receive one free Flea Market ad (subject to our editing) on a space available basis only. Repeat insertions of hamfest ads pay the non-commercial rate.

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DEADLINE 15th of second preceding month.

SEND MATERIAL TO: Flea Market, Ham Radio, Greenville, N. H. 03048.

QSLs & RUBBER STAMPS — Top Quality! Card Samples and Stamp info — 50¢ — Ebbert Graphics 5R, Box 70, Westerville, Ohio 43081.

4D32 Transmitting tubes new at $30.00 each. Limited supply. Collection of 30 years receiving tubes only one dollar each. Send list of needs for availability. Trade commercial test equipment for machinist tools from retired machinists. W5QJT, P.O. Box 13151, El Paso, TX 79913 (915) 581-2017.

WANTED: Old RCA, Western Electric, tubes, speakers, amps. (713) 728-4343, Corb, 11122 Atwell, Houston, Texas 77096.

NEED: Service manual/schematic for HQ-110. I'll copy and return or you send good copy. State price. N3ALA.

TRAVEL-PAK QSL KIT — Converts post cards, photos to QSLs. Stamp brings circular. Samco, Box 203-c, Wynantskill, New York 12198.

WANTED: *Operational Heath HW-8 transceiver. Dean Olson*, Box 161, Box Elder, SD 57719

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RADIO ITEMS before 1930 wanted. Buying battery operated radios, horn and cone speakers, radio tubes and parts, radio literature — books, catalogs, magazines, radio advertising signs, posters. Gary Schneider, 6848 Commonwealth Blvd., Parma Heights, Ohio 44130.

DIGITAL AUTOMATIC DISPLAYS for FT-101's TS-520's, and most others. Six 1/2" digits. Write for information. Grand Systems, P.O. Box 2171, Blaine, Washington 98230. (604) 530-4551.

ELECTRONIC CMOS Keyer kit pcb + parts only $9.95 plus $1.50 shipping. WI res. add 5% tax. Send for free information. BEL-TEK, PO Box 125H, Beloit, WI 53511.

COMPLETE DRAKE "C" LINE: R4-C, T4-C, AC-4, MS-4. Accessories incl. 1.5 kHz and .5 kHz filters, noise blanker, 10 SW Xtals, cables and manuals. All excellent condx. $700. Sony ICF 2001 digital AM/FM/SW portable rcvr. $175. Mark, WB9AQI c/o Milwaukee Sound Studios, 610 N. Water, Ste. 240, Milwaukee, WI 53202. (414) 272-7085 (24 hrs).

RUBBER STAMPS: 3 lines $4.50 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117 SASE brings information.

JOHNSON Matchbox $25, Johnson 6&2 $35, CRT test set $10, Knight tube tester $15, RCA mobile FM test set $20. K6KZT, 2255 Alexander, Los Osos, CA 93402.

"THE SWAP LIST" has bargains galore. Subscribe now! 6 months for $4.00; 1 year only $6.50. The Swap List, Box 988-H, Evergreen, CO 80439.

ENGINEERING SOFTWARE — CP/M, MSDOS, TRSDOS, free flyer. PLOTPRO graph printing program. Linear/log/semilog, multiple plots, grid lines, labeling $52.95. ACNAP — analyzes active/passive electronic circuits. Monte-Carlo Worst Case, sensitivities $52.95. SPP — Signal processing, FFT, linear/nonlinear, LaPlace, transient analysis, $62.95. BV Engineering, Box 3429, Riverside, CA 92519. (714) 781-0252.

HAM RADIO MAGAZINE Collection: Bound volumes 1969, 1970, 1971, 1972, 1973, 1974, 1975. In HR binders 1976, 1977, 1978, 1979. Loose copies 1980, 1981, 1982, 1983. Best offer for lot. W4UCH, P.O. Box 1065, Chautauqua, NY 14722. (716) 753-2654.

REPAIR, ALIGNMENT, calibration. Collins written estimates $25; non Collins $50. K1MAN. (207) 495-2215.

FOR SALE: Yaesu FT DX400 transceiver. Has no final tune-up. $150.00 shipped. Phone (518) 883-5007.

CHASSIS and cabinet kits. SASE K3IWK.

ELECTRON TUBES: Receiving, transmitting, microwave . . . all types available. Large stock. Next day delivery most cases. Daily Electronics, 14126 Willow Lane, Westminster, CA 92683. (714) 894-1368.

WANTED: Cash paid for used speed radar equipment. Write or call: Brian R. Esterman, PO Box 8141, Northfield, Illinois 60093. (312) 251-8901.

WANTED: Old microphones, remote mixers other mic related items. All pre-1935. Box Paquette, 107 E. National Avenue, Milwaukee, WI 53204.

STERLING SILVER callsign jewelry pin or tack; $11.95, alligator tie clip; $19.95, all ppd. Info-SASE. Tom's Silver, PO Box 3758, Manchester, NH 03105.

ROHN TOWERS — Wholesale direct to users. All products available. Write or call for price list. Also we are wholesale distributors for Antenna Specialists, Regency, and Hy-Gain. Hill Radio, P.O. Box 1405, 2503 G.E. Road, Bloomington, IL 61701-0887. (309) 663-2141.

RECONDITIONED TEST EQUIPMENT $1.00 for catalog. Walter, 2697 Nickel, San Pablo, CA 94806.

NH ALL BAND QTH. South slope, Crotched Mountain, Greenfield. Lodge and guest house at 920' elevation. 1200' from highway and power mains. 59 wooded acres, elegant privacy, paved access $95,000.00. Clarence ("CW") Farr, Broker (603) 547-2053.

FOX-TANGO Newsletters — Since 1972, the prime source of modifications, improvements, and repair of Yaesu gear, free to Club members. Calendar year dues still only $8 U.S., $9 Canada, $12 elsewhere. Includes five year cumulative index by model numbers, or send $1 for index and sample Newsletter. Fox Tango Club, Box 15944, W. Palm Beach, FL 33416.

RTTY-EXCLUSIVELY for the Amateur Teleprinter. One year $7.00. Beginners RTTY Handbook $8.00 includes journal index. P.O. Box RY, Cardiff, CA 92007.

IMRA International Mission Radio Assn. helps missioners — equipment loaned; weekday net, 14.280 MHz, 2-3 PM Eastern. Br. Frey, 1 Pryer Manor Rd., Larchmont, NY 10538.

"HAMS FOR CHRIST." Reach other Hams with a gospel tract sure to please. Clyde Stanfield, WA6HEG, 1570 N. Albright, Upland, CA 91786.

TENNATEST — Antenna noise bridge — out-performs others, accurate, costs less, satisfaction guaranteed, $41.00. Send stamp for details, W8URR, 1025 Wildwood Road, Quincy, MI 49082.

WANTED: Early Hallicrafter "Skyriders" and "Super Skyriders" with silver panels, also "Skyrider Commercial", early transmitters such as HT-1, HT-2, HT-8, and other Hallicrafter gear, parts, accessories, manuals. Chuck Dachis, WD5EOG, The Hallicrafter Collector, 4500 Russell Drive, Austin, Texas 78745.

HAVE A-M CAPABILITY? Join S.P.A.M. (Society for Promotion A-M). Membership is free. Write S.P.A.M. c/o F. Dunlap, 14113 Stoneshire, Houston, TX 77060.

VERY in-ter-est-ing! Next 4 issues $2. Ham Trader "Yellow Sheets", POB356, Wheaton, IL 60189.

QUALITY COMPONENTS Transco #11300 UHF relays dual 28VDC coils milspec new $80.00. FXR UHF relay 26VDC coil hermetic milspec new $45.00. Collins filter F455FA15 $65.00. High voltage diode stacks 15kV one ampere milspec new $9.25. Power solenoids 3PST 120A new $45.00; 50A new $35.00 28VDC coils. Carborundum #887 100 ohm 50 watt noninductive resistors new $9.75. Prices prepaid insured. W6FR (714) 871-3607.

NEW 1.2 GHz antennas from NCG. Base or repeater use: (includes mounting hardware) #1260 - 17 step, 1/2 wave collinear, 10.2 dBi gain, VSWR less than 1.5-1, 100 watts, 50 ohm coax. $145.00. #1217 - 17 step, 1/2 wave collinear, 10.8 dBi gain VSWR less than 1.5-1, 30 watts, 50 ohm coax. $85.00. Mobile antenna: #1270M - 10 step, 1/2 wave collinear, 8.5 dBi gain, VSWR less than 1.7-1, 20 watts, type "N" connector $54.50. For more information contact: NCG, 1275 North Grove St., Anaheim, CA 92806.

ATTENTION KENWOOD and Icom owners. Informative separate newsletters. 5th year of publication. Back issues available. Add more selectivity to your 430, 830, and 930. Magicom RF clipper processor for TS-120, 130, 430, 520 and 820. Call for prices. IC-730 FM kit $79.95. TS-830S FM kit $94.95. Send SASE (37¢) for free brochure to International Radio, Inc., 364 Kilpatrick Ave., Port St. Lucie, FL 33452 (305) 335-5545. Master/Visa accepted.

LIKE TO BUY, sell or trade? For the best deals, subscribe to the Ham Boneyard. A complete stockyard of radio and/or related electronic equipment. Second year, published twice monthly, 24 issues annually. Send SASE for free issue. 364 Kilpatrick Ave., Port St. Lucie, FL 33452.

FOR RENT: "Ham-Home", Northern Virginia, 5 miles southwest of Pentagon, 3 bedrooms, family room, fifty foot telephone pole, tribander, 14 el. two-meter, 40/80 meter dipole, 1/2 acre, fenced yard. Available August '84. K1CTK/4. Use callbook address. Phone 703-379-7437 evenings

MARCONI WIRELESS TELEGRAPH CO. OF AMERICA. We have a small quantity of ORIGINAL STOCK CERTIFICATES, signed by the Vice-President and Treasurer of the Company, sealed, issued, and cancelled. MARCONI first proved that messages actually could be sent by wireless, and sparked the imagination of those who became the first HAM RADIO OPERATORS. These certificates, dated in the 1910's, are colorful with an exceptionally nice VIGNETTE of Miss Liberty sitting between globes of the East and West hemispheres, with transmission towers on each side. Each comes beautifully framed and matted, including glass protection, ready to add to the decor of any room or shack. ORIGINAL MARCONI CERTIFICATES cost much less than limited edition art works, represent the beginning of our great hobby, may increase in value (Scripophily, the collection of old stock and bond certificates, is in its infancy and becoming more popular), represents the beginning of a giant corporation (RCA was formed with the express purpose of buying out the Marconi Wireless Telegraph Co. of America), makes a unique addition to any home, and makes a great gift. Ten day unconditional money back guarantee. Price complete and delivered (we pay shipping charges): $49.95 (Pa. residents must add $3 for sales tax). Send check (personal checks — O.K. Place your call on your check and we will ship immediately without waiting for the check to clear your bank), Money Order, or Master Card/VISA information to: AUTHENTIC CERTIFICATES, RD 2073, Fleetwood, Pa. 19522. Phone: 215-944-0627. Checks received after the MARCONI Certificates are sold out will be returned uncashed.

Coming Events

ACTIVITIES

"Places to go..."

ILLINOIS: The Quad-Co. ARC will sponsor the 27th annual "Breakfast Club" Hamfest, July 21 and 22, Terry Park, Palmyra. All other groups are invited. Please give prior notice to Hamfest Committee. Dancing and movies Saturday night. Bring a basket lunch. Sandwiches and soft drinks available on grounds. Games, contests, golfing and fishing. Bring your swap gear. Camping facilities open from Friday afternoon to Monday morning. Talk in on 3973 kHz. Write: Hamfest, c/o Quad-Co. ARC, 602-D East Walnut, Chatham, IL 62629.

SOUTH CAROLINA: The Charleston Amateur Radio Society will host the annual Charleston Hamfest, July 14 and 15, Omar Shrine Temple. Commercial tables $40.00. (over 2, $20.00 each additional). Flea market tables $5.00 each. Talk in on 146.19/79. For information/table reservations: Hamfest Committee, P.O. Box 70341, Charleston Heights, SC 29405.

TEXAS: The Austin ARC and the Austin Repeater Org. will sponsor Summerfest '84, August 10, 11 and 12, Austin Marriott Hotel, I-35 at Hwy. 290. Admission $5 advance (deadline July 31). $7.00 at the door. Indoor swapfest, dealers' showcase, transmitter hunt. Featured speakers: KG5U, president of the NASA/Johnson Space Center ARC, KO5I of AMSAT, K5FOG and WA5MWD on packet radio. Ladies' programs. Reserved swapfest tables $1.00 each, limit two. Talk in on 146.34/.94. For more information: Austin Summerfest '84, P.O. Box 13473, Austin, TX 78711.

WEST VIRGINIA: The 6th annual TSRAC Hamfest, Sunday, July 22 at Wheeling Park. 9 AM to 4 PM. Admission $3.00 by reservation. Dealers, exhibits all under roof. Refreshments, family park activities, flea market, auctions, free parking. All spaces free both dealers and flea market. For brochure, information and map contact: TSRAC, Box 240, RD 1, Adena, OH 43901. (614) 546-3930.

TENNESSEE: The Radio Amateur Transmitting Society is pleased to announce the 6th annual Nashville Ham and Computer Fest, July 29, at the Nashville Municipal Auditorium. No admission charge. Tables availble for $5.00. For further information: SASE to Willie Porter, KB4BLL, 4907 Idaho Avenue, Nashville, TN 37209.

NEW YORK: The Allied Armored Force Amateur Radio Nationwide Emergency Team, (AFAR NET) will hold an eyeball bivouac, July 28, Sheraton Inn in Canandaigua starting at 1 PM. All Amateur Radio operators who have been assigned or attached to an armored unit are invited to attend the get-together. For additional information: Harry B. Thomsen, W2PJH, 348 Jefferson Avenue, Apt. 15, Canandaigua, NY 14424. Please SASE.

NEW JERSEY: Sussex County ARC will sponsor "SCARC '84", Saturday, July 14, at the Sussex County Fairgrounds, Plains Rd., Augusta. Doors open at 8 AM. Admission $2. Indoor tables $5 advance, $6 at door. Tailgate space $4 advance, $5 at gate. Food and refreshments. Free parking. Talk in on 90/30 and 52 simplex. For further information: Donald R. Stickle, K2OX, Weldon Rd., RD #4, Lake Hopatcong, NJ 07849. (201) 663-0677.

WISCONSIN: The Eau Claire Amateur Radio Club's annual Hamfest, Saturday, July 14, 4-H Buildings in Eau Claire. 8 AM to 4 PM. Tickets $2.00 in advance, $3.00 at door. Free tables and coffee. Talk in on 31/91 and 52 simplex. For information, tickets SASE to Gene Lieberg, KA9DWH, 2840 Saturn Avenue, Eau Claire, WI 54703.

FLORIDA: The 11th annual Greater Jacksonville Hamfest, August 4 and 5, Orange Park Kennel Club, US 17 south near I-295. 8 AM to 5 PM Saturday and 9 AM to 3 PM Sunday. Registraion $4.00. Swap tables $9/one day, $15/weekend. Exhibitors, large swap table area, forums and programs. Special discounts and promotions available to exhibitors contracting for space prior to July 15. Plenty of free parking. For information, swap tables, registration: Mike Parnin, N4EPD, 6716 Diane Rd., Jacksonville, FL 32211.

MICHIGAN: The Copper Country Radio Amateur Association is hosting the 1984 Upper Peninsula Hamfest, July 28, Memorial Union Cafeteria, Michigan Technological University campus, Houghton. For further information contact: Howard Junkin, N8FHF, Co-chairman UP Hamfest, 106 West South St., Houghton, MI 49931. (906) 482-4630.

MONTANA: The Great Falls Area ARC's 50th annual Glacier-Waterton International Hamfest, July 20, 21 and 22 at Three Forks Campground on the southern edge of Glacier National Park. QCWA meeting and contests, 2m transmitter hunts. Preregistration is $8.50 and includes Saturday night dinner (bring your own meat and utensils) and Sunday morning breakfast. Talk in on 52 and 34/94. For more information SASE to Shirley Smith, KC7OA, 1822 - 14th Avenue South, Great Falls, MT 59405.

MARYLAND: The Baltimore Radio Amateur Television Society again presents the BRATS Hamfest and Computerfest, Sunday, July 29, Howard County Fairgrounds, Route 144 at Route 32, West Friendship. Indoor tables with A/C $20.00 each. Indoor tables no A/C $10.00 each. Outdoor tailgating. RV hookups available. For table reservations and information: BRATS, P.O. Box 5915, Baltimore, MD 21208 or call Mayer Zimmerman, W3GXK (301) 655-7812. Talk in on 146.76 (−600), 147.03 (+600) or 52 simplex.

WEST VIRGINIA: Wheeling Hamfest, Sunday, July 22, Wheeling Park. Flea market, auction, dealers welcome. Under roof tables available. Admission $3.00. For information/reservations: TSRAC, Box 240, RD 1, Adena, OH 43901. (614) 546-3930.

INDIANA: The combined LaPorte-Michigan City Amateur Radio Clubs will sponsor their Summer Hamfest, Sunday, July 15, LaPorte County Fairgrounds, State Road 2, LaPorte. 8 AM to 2 PM. Donation $3.00. Indoor tables by reservation, 40¢/ft. P.O. Box 30, LaPorte, IN 46350.

KENTUCKY: The Central Kentucky ARRL Hamfest, sponsored by the Bluegrass Amateur Radio Society, Sunday, August 12, 8 AM to 5 PM, Scott County High School, Longlick Road and US 25, Georgetown. Tech forums, awards, exhibits. AC facilities. Free outside flea market space. Tickets $3.50 advance and $4.00 at gate. For information or tickets: Edward B. Bono, WA4ONE, P.O. Box 4411, Lexington, KY 40504.

NORTH CAROLINA: The WCARS annual Hamfest, July 28 and 29, Firemen's Training Center, Asheville. Admission $4.00. Preregistration $3.50 before July 15. Talk in 31/91, 16/76 and 52 simplex. For information: Nelda Williams, KA4WPM, P.O. Box 1488, Asheville, NC 28802.

PENNSYLVANIA: The Mid-Atlantic ARC announces its annual Hamfest, Sunday, August 12, 9 AM to 4 PM, rain or shine. Route 309 Drive-In Theater, Montgomeryville. Admission $3.00 plus $2.00 additional for each tailgate space. Bring your own table. Plenty of parking and refreshments. Talk in on WB3JOE/R, 144.66/.06 or 146.52. For information: MARC, P.O. Box 352, Villanova, PA 19085 or call Bob Josuweit, WA3PZO (215) 449-9727.

OKLAHOMA: The Central Oklahoma Radio Amateurs will host "Ham Holiday" and State ARRL Convention, July 21 and 22, Lincoln Plaza Inn & Conference Center, 4445 Lincoln Blvd., Oklahoma City. DX, Packet Radio, satellites, computers with RTTY and a host of special interest events. Unlimited free parking for cars and self-contained RV's. Dealers Saturday and Sunday. Flea Market Saturday only. Pre-registration by July 6, $8.00; $10.00 at the door. Saturday evening banquet $14.00. Sunday QCWA breakfast $7.20. Reserved flea market table $5.00; $8.00 at door (if available). Mail reservations to CORA, P.O. Box 44091, Oklahoma City, OK 73144.

MICHIGAN: ARRL State Convention, sponsored by the Livonia ARC with the help of nine other Radio Clubs. Friday, June 29 and Saturday, June 30, Schoolcraft College campus, 18600 Haggerty Road, Livonia. Friday evening "Eye-Ball Social", 7:30 PM at the Plymouth Hilton with a Wouff Hong initition at Midnight. Convention opens Saturday, 8 AM to 5 PM. Exhibits, forums, seminars, YL activities, operating demos, state nets and more. Convention banquet Saturday evening, 7:30 PM at the Plymouth Hilton. Admission $3.00 at the door; $2.50 advance. Swap tables $1.00/ft. and trunk sales $5.00 space. Talk in on LARC Repeater 145.35 (−600) or the 146.52 simplex. For tables, trunk sales, tickets or information: Neil Coffin, WA8GWL, Livonia ARC, P.O. Box 2111, Livonia, MI 48151.

NEW HAMPSHIRE: Fly-in to the 2nd largest Amateur Radio/electronic flea market in the state, Saturday, July 14, Manchester Municipal Airport. Sponsored by the New Hampshire FM Association. Starting time 9 AM. General admission $1.00. Sellers $5.00 with own table. Refreshments available. For information: Dick DesRosiers, W1KGZ (603) 668-8880 or Doug Aiken, K1WPM, 30 Meadowglen Dr., Manchester, NH 03103 (603) 622-0831.

OHIO: The Northern Ohio Amateur Radio Society's 7th annual ARRL approved NOARSFEST, July 21, Lorain County Fairgrounds, Wellington. 8 AM to 4 PM. Huge black-topped flea market area $1.00 per car space. Setup 6 to 8 AM. Free overnight camping Friday night, no hookups. Dealers indoor 8' tables $8.00 each. Donation $3.00 advance, $3.50 gate. Children under 12 free. Talk in on 144.55 - 145.15. Mobile checkin 146.52. For advance registration mail check to John Paul Jones, WA8CAE, 4612 Timberview Drive, Lorain, OH 44052 (216) 282-4256.

NORTH CAROLINA: The Cary ARC's 12th annual Mid-Summer Swapfest, Saturday, July 21, 9 AM to 3 PM, Lion's Club Shelter, Cary (near Cary High School). Open auction starting at 1:00 PM. We sell anything! No commission. No admission. Talk in on 146.28/.88; 147.75/.15; 146.52/.52. For information: Cary ARC, P.O. Box 53, Cary, NC 27511.

NEW JERSEY: The West Jersey Radio Amateurs' 6th annual Hamfest, July 15 (rain or shine) 9 AM to 3 PM, Super 130 Drive-in Theatre, Route 130, Edgewater Park. Registration $3.00. Tailgating $3.00 additional with own table. Vendors' only setup starting at 7 AM. For information or tickets SASE to: Mary Lou Shontz, N2CLX, 107 Spruce Lane, Route 16, Mount Holly, NJ 08060. (609) 267-3063.

MISSOURI: The 22nd annual Zero-Beaters ARC Hamfest, July 15, 9 AM to 3 PM, Washington, MO, Fairgrounds. Free admission. Free flea market area. Limited rental spaces under pavillion. Advance reservation advised. Cake walk, candy scramble, traders row, refreshments and food available. Talk in on 147.24-84, 146.52. For information: Zero-Beaters ARC, Box 24, Dutzow, MO 63342.

COLORADO: The Ski Country ARC's third annual Swapfest and CCARC meeting, July 28, 9 AM to 3:30 PM, CMC Building, 1402 Blake, Glenwood Springs. Tables $5.00 each. Half table $3.00. Speakers, demos. Talk in on 146.07/67. For information: Bob Ludtke, K9MWM, 1001 Grand Ave., Glenwood Springs, CO 81601. (303) 945-5966.

VERMONT: The annual BARC International Hamfest, Saturday, August 11 and Sunday, August 12, Old Lantern Camp Grounds, Charlotte. Tickets $4.00 for both days. Heterodynes under 12 free. Outdoor flea market space $2.00; indoors $5.00. Overnight camping available. Talk in on 34/94, 01/61, and 52 simplex. For information: Roger, WA1OZE, Burlington ARC President, P.O. Box 312, Burlington, VT 05402.

WASHINGTON: W7DK, the Radio Club of Tacoma, presents Hamfair '84, August 11 and 12, Olsen Auditorium, Pacific Lutheran University campus. Registration $5.00. Trailer and dormitory space available on campus at reasonable rates. Saturday night banquet. Commercial space and flea market tables available by advance registration. Talk in on W7DK repeater, 147.88/28. For information or registration: Grace Teitzel, AD7S, 701 South 120th, Tacoma, WA 98444.

IOWA: The 18th annual Central States VHF Conference, July 27-29, Sheraton Inn, Cedar Rapids. Technical presentations, DXpedition reports and organizational meetings. Plan to attend and meet your friends. For more information: Barry Buelow, WAØRJT, 4110 Emerson Ave N.E., Cedar Rapids, IA 52402.

ILLINOIS: The DuPage Amateur Radio Club Hamfest/Computerfest will be held on Sunday, July 8, 1984, at the Downers Grove American Legion Post 80. Large outdoor flea market and swappers row. Indoor commercial exhibits available. General admission $2.00, $3.00 for swappers row sellers. Tickets available at the gate only. Plenty of free parking space. Food and drink available. Talk-in on 145.490 Rpt. or 223.74 Rpt. For more information, send SASE to W9DUP, P.O. Box 71, Clarendon Hills, IL 60514 or call 312-971-3298 between 8 A.M. and 9 P.M.

OPERATING EVENTS

"Things to do..."

JULY 21: The Tank-Automotive Command ARC will operate W8JPW from 1300 to 2000Z to commemorate the 43rd year of the Detroit Arsenal. Frequencies: Phone 7.274, 21.400 and 146.49 MHz. CW 7.055 from 1500-1700Z. Send 9 × 12 SASE with QSO number and frequency for certificate to W8JPW, U.S. Army Communications Command, Att: CCNC-TAC-M, 28251 Van Dyke, Warren, MI 48090.

SEPTEMBER 9-15: The Soutern Counties Amateur Radio Association (SCARA) is planning to have a special events station during the Miss America Pageant. Check September Ham Radio for details.

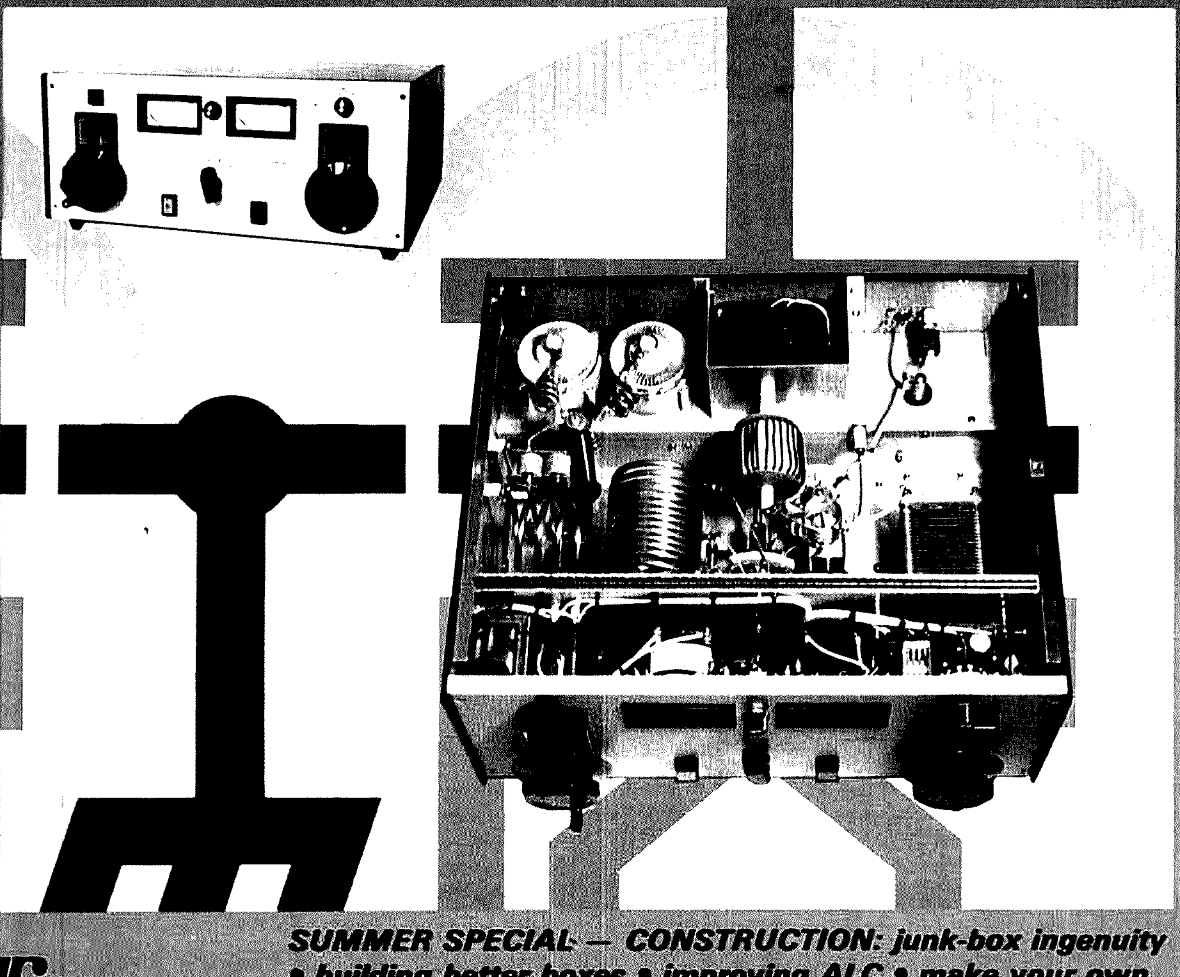
AUGUST 4-5: The Wild Bunch 160 SSB Contest: 0000 GMT Aug. 4 to 2359 Aug. 5. Single operators only. Exchange: RST/State, Country or Province. 10 pts. per QSO. Multipliers: total number of states, countries, VEs. Plaque to winner. Deadline for logs September 6, 1984, to: KA1SR, R J Koziomkowski, 5 Watson Dr., Portsmouth, R.I. 02871.

JULY 14-15: The Eastern Michigan ARC (K8EPV) will commemorate the annual Port Huron to Mackinac Island Yacht Race. 1400Z to 0200Z both days. Frequencies: 3,910, 7,235, or 14,235. Phone 3,710, 7,110 or 21,110 CW. For a certificate send QSL with legal SASE to K8EPV (C.B.A.) or 654 Georgia, Marysville, MI 48040

AUGUST 1984 / $2.50

BUILD A 10 TO 160-METER
LINEAR AMPLIFIER
WITH EIMAC'S NEW 3CX800A7'S

ham radio
magazine



SUMMER SPECIAL — CONSTRUCTION: junk-box ingenuity • building better boxes • improving ALC • make your own air-wound coils • cooling semiconductors • ALSO: using grayline — with or without computer • a 50 MHz Tilton/Greenblum Yagi • W1JR on filters • W6SAI's ham radio techniques

hr
focus on communications technology

ham radio magazine

AUGUST 1984

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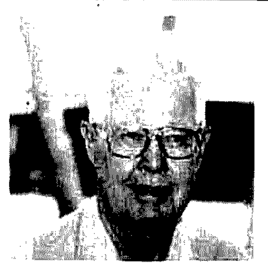


Mason Logan, K4MT

REFLECTIONS

. . . with a little help from our friends

Forrest Gehrke, K2BT

We hate to admit it, but we don't know everything there is to know about Amateur Radio. What's more, the field is so vast, and expanding so quickly, that we can't even *hope* to know it all. We try — but there are times when our individual and collective wisdom fails and we have to turn to others for help. Perhaps the subject at hand is an idea so new that we need to learn more about it ourselves. Or perhaps all we need is confirmation of our opinion that an idea may be unworkable because of some flaw in the concept or method proposed.

When we need a knowledgeable friend to turn to — to review a questionable manuscript, or to clarify, confirm, or correct some detail or theory or practice — we turn to one of the four distinguished members of *ham radio's* editorial board. We look to them for two things: first, for their expertise in their specific areas of knowledge and experience; and second, for their unique perspective on Amateur Radio. Spanning the past and present, with an eye on the future of Amateur Radio, these gentlemen give us a perspective that's easy to lose sight of in the day-to-day business of making sure your copy of *ham radio* arrives each month.

While we'd like to introduce each one of the members of our editorial board to you personally, that's obviously impossible. So we'll do the next best thing and introduce them to you in print.

Ed Wetherhold, W3NQN, first encountered Amateur Radio in 1947 while in training as an Air Force radio technician. A buddy in the barracks had set up a station next to his bunk. Ed was "hooked." He went on to study Radio Engineering at Tri-State University in Angola, Indiana, and graduated in 1956. Since 1962 he's been with Honeywell in Annapolis, Maryland, where he's responsible for testing communications systems. In addition, he distributes surplus 44- and 88-mH toroidal inductors in the United States and in the U.K. to facilitate easy assembly of his filter designs, and writes extensively in the international Amateur press. Ed also serves as technical advisor to the ARRL in the area of his specific interest, passive LC filters.

While Ed found Amateur Radio, Amateur Radio "found" **Forrest Gehrke, K2BT**. At 10, he was building one-tube regenerative broadcast band receivers, using UV-199's and 201A's. One day he inadvertently omitted some turns around the oatmeal box, and some fascinating activity on the 160-meter band spilled into his earphones. First licensed as W9WJD in 1936, Forrest says he was "bitten by the DX bug shortly afterward when, during sunspot peak openings, you could work the world on 10 meters with no discernible power output." He earned an E.E. degree from the University of Wisconsin and spent 30 years trying to "push the state of the art" in electron tubes. By the early 1960's, he was working in the development of a 10 GHz solid-state power source for the MoonLander's retro-rocket landing control radar. He now works in the data communications industry, developing packet networks that communicate at 56K bits per second.

World War II, marriage, and sharing in the raising of eight children caused what K2BT calls "slight intermission" in Amateur Radio activities until 1968, when he resumed the push for DXCC he'd begun in 1939. He reached 5BDXCC in 1974, Honor Roll in 1978, and is currently aiming for Honor Roll Status solely in the 80-meter band. His special interest: antennas.

Mason Logan, K4MT, studied physics at CalTech and Columbia and worked with Bell Telephone Laboratories for 50 years. His early work was in telephone transmission and signalling research; later he designed circuits for underwater acoustic and magnetic proximity fuzes, high-loop gain negative feedback amplifiers, and servomechanism circuits for the development model of the NIKE guided missile analog computer.

Now and then a problem will come along that demands a theoretical solution. If it pertains to any aspect of electromagnetics, we turn to **Bob Lewis, W2EBS**. Bob started out in Amateur Radio in 1921 (as did K4MT) when he was a 13-year old Boy Scout leafing through the pages of his handbook in search of another merit badge to win. The "Wireless" badge caught his eye and he ran to the store for some oatmeal — not for the cereal, or course, but for the box that it came in. Finally licensed in the early 1930's, he attended Virginia Military Institute and graduated from the University of Pennsylvania. Graduate study — including an M.B.A. — followed.

Bob worked in research at RCA from 1932 through 1939, when he left to join the research team at CBS. (While at RCA, incidentally, he worked in the design of the television transmitting antennas atop the Chrysler Building in New York City.) He served overseas during World War II, working in radar countermeasures, with the office of Scientific Research and Development. After the war he returned to CBS and then went to Federal Telegraph and Radio. After ITT bought Federal, he joined the ITT research group in Nutley, New Jersey. He joined a small group of friends and colleagues in Prodelin, Inc., and spent the following 20+ years, until retirement, working in the development of transmission lines and antennas.

Over the coming months we'll be looking to expand the editorial review board to include others of similar standing, but with different areas of expertise. If you have an area of special interest and achievement in Amateur Radio — and aren't too modest to admit it — get in touch with us. You don't have to be one of the Founding Fathers of Amateur Radio; if you "know your stuff," and want to play a meaningful role in the production of a high-quality technical publication, let us know. *We'd like to know you.*

Dorothy Rosa, KA1LBO
Assistant Editor

presstop de W9JUV

SOME COMMERCIAL INCURSION INTO THE AMATEUR 220-225 MHZ BAND is still very possible, according to informed observers of the Washington and Land Mobile scene. Despite strong denials at high FCC levels, there's an ongoing belief in some circles that at least a portion of the band (some say as much as half!) could be reallocated to Land Mobile in the near future. If so, it would likely go for a narrowband technology such as ACSB.

A Proposal To Permit Novice Phone Operation On 220 MHz has been submitted to the FCC by WA2MCT and WD5DON as a Petition for Rulemaking. Opposition to their proposal since it was first suggested has been very strong, both at the Dayton Hamvention 220 MHz Forum and in letters to "220 Notes," K9XI's national 220 MHz Newsletter.

In An Unrelated Move, A 224.750 MHz Experimental License has been granted to the University of Illinois' Wallops Island, Virginia, test facility "to support research in ionospheric radio propagation required by U.S. Government contract."

420-430 MHZ IS NO LONGER AVAILABLE TO U.S. AMATEURS located within 75 miles of the Canadian border. The ban results from Canada's decision several years ago to allocate the bottom 10 MHz of the 70-cm band to land mobile, and a consequent agreement between its DOC and the FCC to protect Canadian land mobile users from possible U.S. Amateur interference. Though the protection band, which actually exceeds 75 miles in some areas, has theoretically been in effect for some time, the FCC has not yet begun to actively enforce it.

Included In The Protection Band Are Such Major U.S. Cities as Seattle and Duluth, most of Michigan (including Detroit), Toledo, Cleveland, Erie, at least half of the states of New York, Vermont, and New Hampshire, and most of Maine, including Bangor! A 75-mile band of protection also extends along the Alaska-Canada border, encompassing Juneau and Ketchikan.

U.S. Land Mobile Stations Using Frequencies Between 30 And 470 MHz In The Protected Areas are also affected, but can receive clearance for licensing on a non-interfering basis. U.S. land mobile stations with an ERP under five watts do not require any Canadian coordination. However, there is no such leeway for Amateur operations along the borders.

Unusual Signals May Be Encountered In The 70-cm Band By Amateurs in much of New York, New Jersey, and Pennsylvania in coming months. Grumman Aerospace has received an FCC experimental license to operate on various frequencies between 424 and 446 MHz within a 150-mile radius of Binghamton, New York, in connection with work on the E-2C system.

A THIRD AMATEUR HAS BEEN NAMED TO BECOME AN ASTRONAUT BY NASA. Ron Parise, WA4SIR, a scientist employed by NASA at Greenbelt, Maryland, is scheduled to make his first trip on the Space Shuttle in 1986. He joins W5LFL, who conducted the first Amateur operation from space late last year, and WØORE, who's up for his first Shuttle trip next spring.

Amateur Operation During WØORE's Upcoming Shuttle Flight has been formally proposed by the ARRL and AMSAT. In their joint proposal the two groups stated their goal was to involve as many Amateurs as possible, particularly through school and club stations. In addition to 2-meter FM such as W5LFL used, a 10-meter downlink for 2-meter audio and SSTV pictures from the Shuttle has also been suggested. NASA's decision is expected soon.

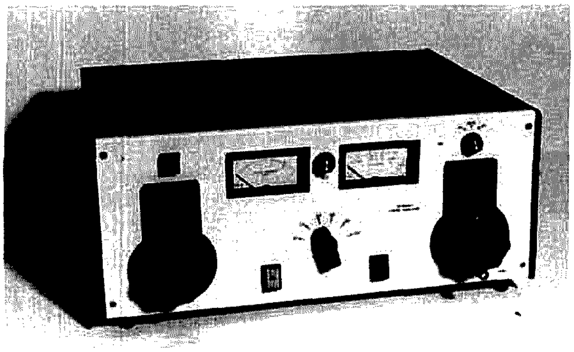
SIMPLEX AUTOPATCHES MUST HAVE A SEPARATE CONTROL MEANS ON ANY AMATEUR BAND, not just below 220.5 MHz. Confusion arose with the ARRL Executive Committee's recent adoption of a requirement that all QST ads for simplex autopatch devices state their use is not permitted "...in the 2-meter band, or on any other frequency below 220.5 MHz..." without a separate control link. The FCC requires positive transmitter control on all Amateur frequencies—not just 2 meters—but feels the real issue is whether the simplex autopatch is really "Amateur Radio" rather than how existing rules can be bent to fit its operation.

VOLUNTEER EXAMINER COORDINATORS HAVE NOW BEEN ACCEPTED IN ALL AREAS, with the appointment of W5YI Report Editor/Publisher, Fred Maia, as a VEC for all 13 districts. The Zero District now also has a "resident" VEC, the PHD Radio Club from Missouri; until Maia's acceptance, the First District was the last still lacking a VEC.

ARRL Applied To Become A VEC In All 13 Areas June 27. In their 1-1/2 page proposal, down sharply from an earlier 70-80 page draft, the League still remained adamant that its active participation wouldn't begin until the FCC authorized collection of exam fees. However, it still looks very likely that the fee proposal will be adopted in some form before the FCC's summer recess begins August 1. The height of the "Chinese wall" between the League's publishing business and its VEC organization is still a likely problem area.

Well Over 1000 Exams Have Now Been Given Under VEC Direction, and the program seems to be working quite well in most areas. Most VECs expect to be ready to give Advanced and Extra Class exams by the end of July, and the Second District VEC, Metroplex, will be giving Novice through Extra Exams at the ARRL National Convention in New York July 21-22.

ARRL'S PETITION TO HAVE CABLE TV KEPT OFF THE AMATEUR BANDS has been rejected by the FCC. However, in their rejection the Commissioners put the cable TV industry firmly on notice that it has an obligation to prevent and remedy leakage problems, to all services.



a 3CX800A7 linear amplifier

1500 watts out on 10-160 meters, using a pair of EIMAC's new compact tubes

It took me about thirty seconds to accept *ham radio's* invitation to design and build a new high-frequency linear amplifier centered around EIMAC's new 3CX800A7 triode. I decided immediately that the design should capitalize on the small size of the tubes, cover the 10 through 160 meter bands, and be capable of meeting the new 1500 watt output power limit on a continuous basis. In other words, I wanted a small desk-top amplifier that "growled." My approach to the project was similar to the modular technique described in my previous *ham radio* article[1] and to the structured approach to equipment design outlined earlier in *QST*.[2]

In this article, I have made a special effort to provide the detail necessary to allow readers to easily duplicate my efforts. I have provided detailed layouts of the circuit boards and tried to identify sources for the parts where available. Readers who duplicate the design exactly should experience few problems.

Some unusual parts — a 500 pF/3000 volt vacuum variable capacitor and an RF vacuum relay — for example — are used. In each case, effort has been made to identify these parts and suggest alternative components that are more readily available.

I have also included construction details for a matching high voltage power supply. Remember that a well-regulated power supply is one key to making an amplifier perform well.

RF deck circuit design

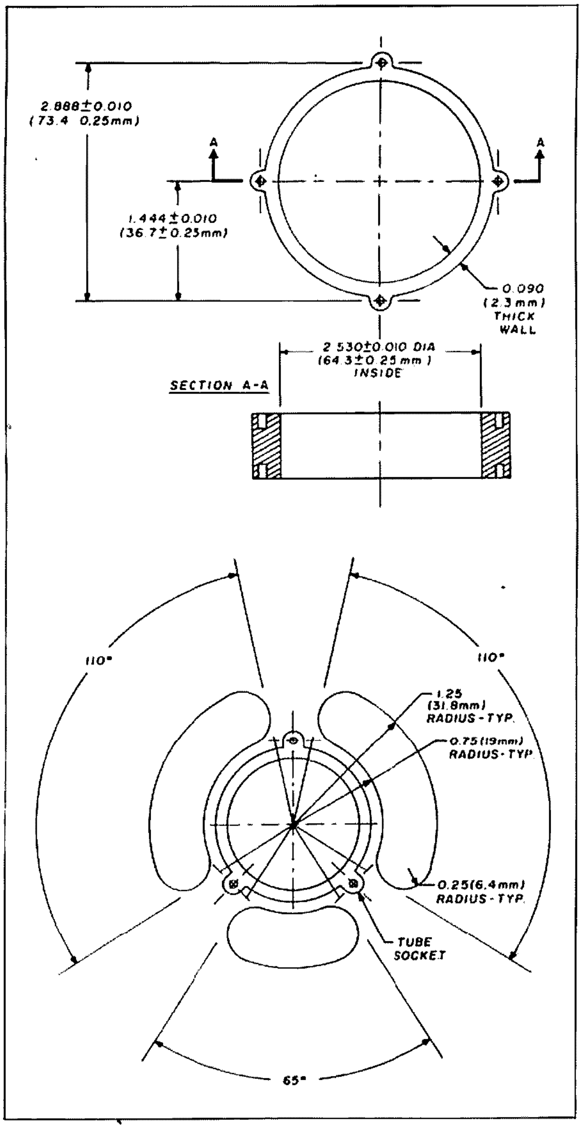
The amplifier is designed around a pair of EIMAC 3CX800A7 high-mu power triodes that are capable of 15 dB gain. The tubes are very compact, measuring only 2-1/2 inches (6.35 cm) high and 2-1/2 inches in diameter. A matching socket, chimney, and plate connector are also available.

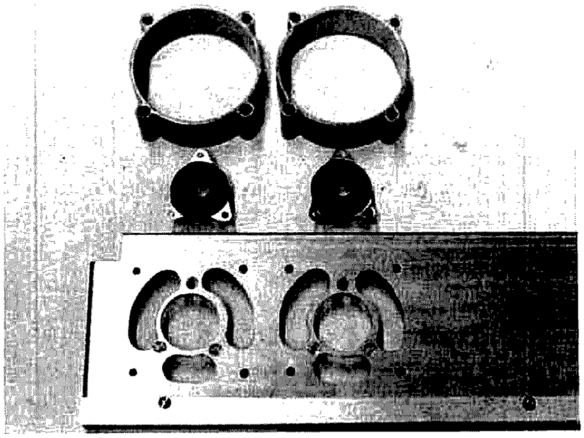
The amplifier uses a tuned input network (**fig. 1**) ganged to the main band switch to minimize distortion products and provide a good match to the exciter. Maximum SWR presented to the exciter is 1.3:1. Because of the high gain of the tubes (i.e. 15 dB) only about 60 watts is needed to drive the amplifier to the new 1500 watt output power limit. Therefore, a very effective ALC circuit (see **fig. 2**) also has been included to control drive power. (See ALC module PC board and component layout.)

Output matching is accomplished using a Pi-L circuit designed for a *Q* of 12. Toroids in both the input and output circuit have been used for compactness.

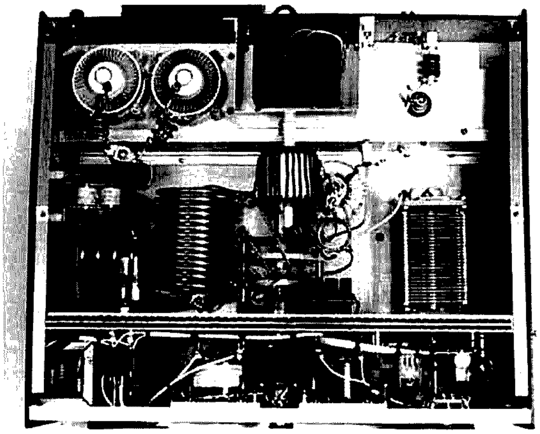
Also included is a very effective grid trip circuit that latches the amplifier out of operation should the grid current exceed 90 milliamperes (45 milliamperes per tube). This feature provides positive protection for the tubes against excessive grid current. This is especially important with these tubes since the grids are capable of dissipating only 4.0 watts. If the protective circuit does trip, it can be reset by pushing a front panel mounted switch. However, the reason it tripped

By Jerry L. Pittenger, K8RA, 2165 Sumac Loop South, Columbus, Ohio 43229





Mounting details for 3CX800A7. Sockets and chimneys available from EIMAC.



Top view of amplifier: Note shield for filament transformer; finger stock isolates meter compartment from main RF compartment.

in the first place should be determined before resetting the switch.

Additional protective devices include a solid-state time delay circuit that prevents operation of the amplifier for approximately 3 minutes until the indirectly heated cathodes of the tubes are properly conditioned. The HV can not be turned on during this period. After warm-up, a relay actuates that allows the amplifier to be keyed up by the exciter and also sends AC power to the HV power supply to allow HV to be applied to the plates of the tubes.

A regulated 26 VDC power supply has been included to provide power for meter and switch pilot lights and relay control. The supply is regulated to avoid pilot light dimming under varying current loads.

amplifier control circuitry

The control circuitry for this amplifier is quite simple. 110 VAC enters the RF deck control circuit (**fig. 3**) from the HV power supply. When the front panel FIL power switch is actuated, AC power is applied to the blower, filament transformer and 26 VDC regulated power supply module. (See PC board and component layout.) The 26 VDC supply, in turn, applies power to the timing module, which starts the three minute warm-up cycle. When the warm-up cycle is over, the meter lights come on indicating that the amplifier is ready to operate. Note, actuating the front panel "HV" power switch will not send power to the HV power supply until the 3-minute warm-up cycle has completed.

The 26 VDC power supply uses an LM317 voltage regulator which can malfunction in the presence of strong RF fields. Consequently good design practices require careful placement. In this design, the 26 VDC supply is located up in the front meter compartment far away from RF.

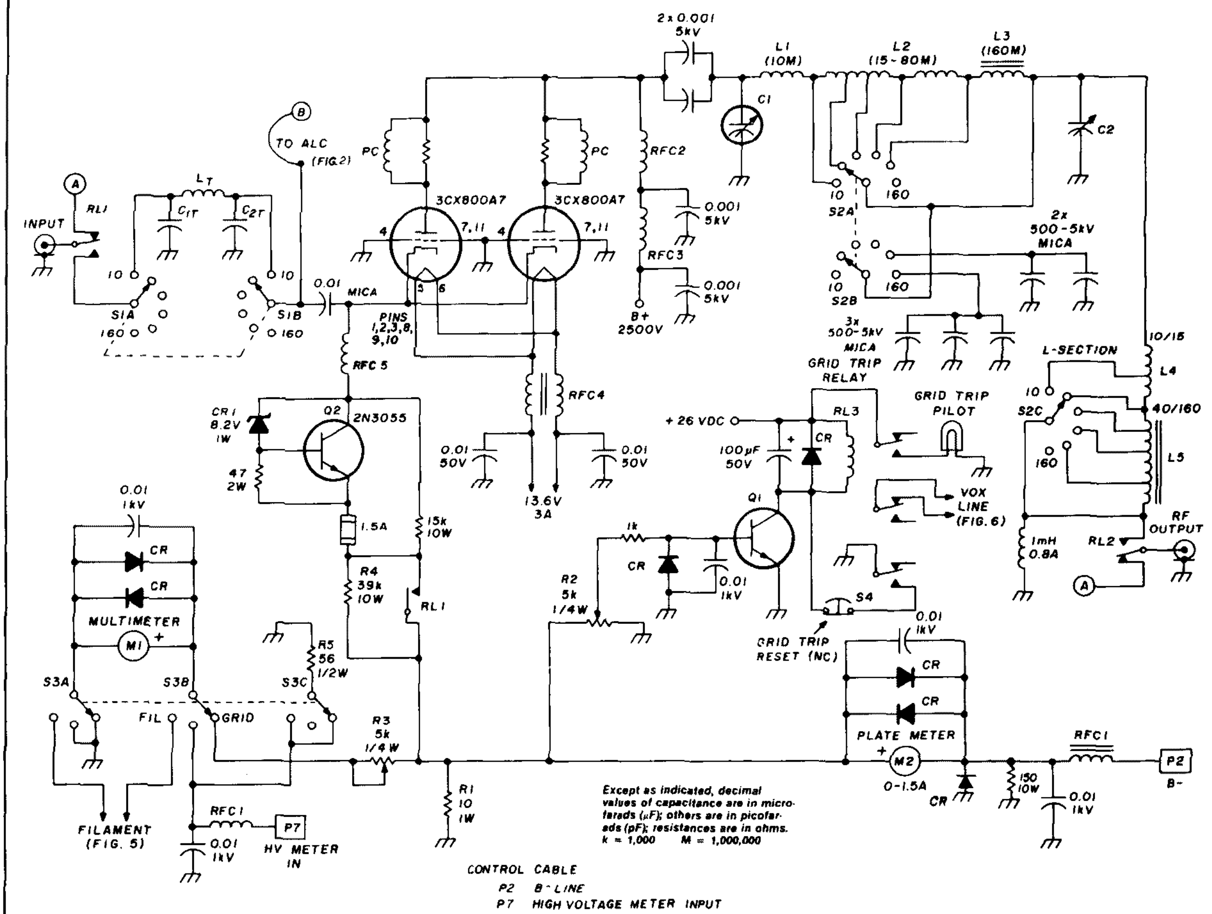


fig. 1. 3CX800A7 RF deck schematic.

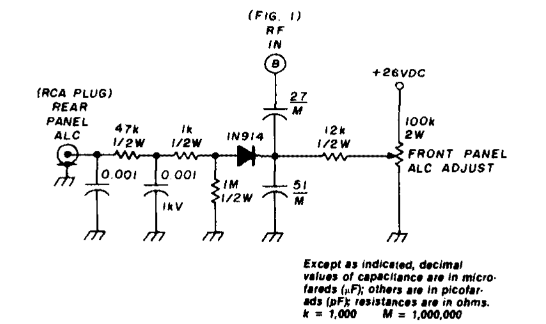
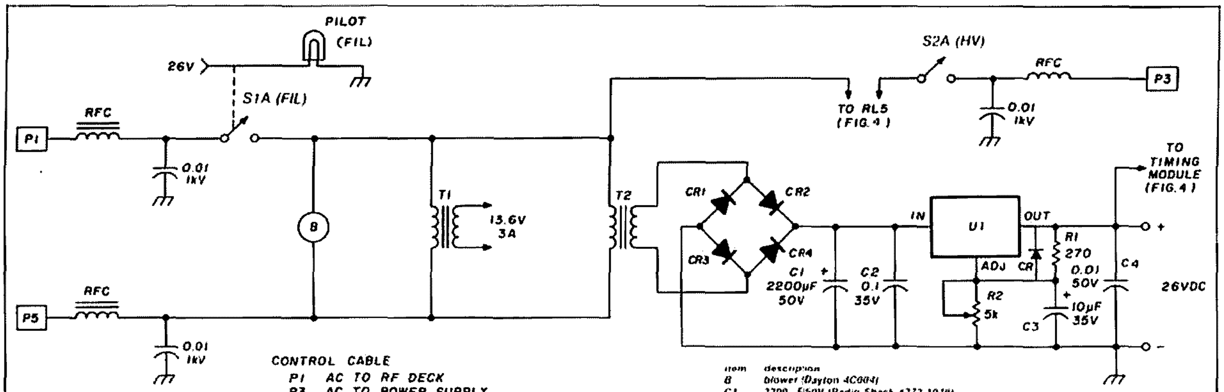


fig. 2. ALC module schematic.

grid-trip protection circuit

A grid trip module has been included to disable the amplifier should the grid current rise to levels dangerous to the tubes. A pair of 3CX800A7s is rated for a maximum grid current of 120 milliamperes (mils). Under normal operation, a pair of tubes draws about 30 mils. Therefore, I have the circuit adjusted to trip at 90 mils of grid current. When the circuit trips, relay RL3 (see **fig. 1**) is actuated which breaks the VOX line, *thus deactivating the amplifier and lighting the "grid* trip reset" push button on the front panel of the amplifier. It is necessary to push the RESET button to put the amplifier back into operation. Of course, one should determine why the amplifier exceeded 90 mils before resetting.

The circuit is actually quite simple. As grid current flows through R1, a voltage is developed across the resistor. The grid trip actuates when transistor Q1 is *turned on (approximately 0.6 volts* appears on the base). For example, if the base of Q1 were tied directly to R1, 60 milliamperes would generate 0.6 volt ($E = IR = 0.060 \times 10 = 0.6$ *volts*). But this current is too low in normal operation to trip the grid current protection circuit. Therefore, a voltage divider is needed if a current different than 60 milliamperes is desired to actuate the grid trip. This voltage divider



CONTROL CABLE
P1 AC TO RF DECK
P3 AC TO POWER SUPPLY
P5 AC COMMON

Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

item	description
B	blower (Dayton 4C004)
C1	2200 µF/50V (Radio Shack #272-1040)
C2 4	0.1 µF tantalum (Radio Shack #272-1432)
C3	10 µF/35V (Radio Shack #272-1025)
CR	bridge diode 50 PIV/1.4A (Radio Shack #276-1151)
IC	LM317 adjustable 1.5A regulator (Radio Shack #276-1778)
RFC	10 turns No. 14 enamel wire on 1/4"-diameter ferrite rod
S1 S2	rectangular pushbutton switch (2 N.O. and 2 N.C.) (Newark) Alcoswitch 16TL5-22 Alcoswitch 61-4 yellow lens Alcoswitch 61-2 red lens
T1	13.6VAC/3A custom transformer (Peter Dahl)
T2	24VAC/1A Triad F46-X (Newark) (Radio Shack #273-1480)

fig. 3. Control circuit, 26 volt DC power supply.

parts list for fig. 1

item	description
C1	500 pF/3kV vacuum variable (air variable can be substituted)
C2	850 pF/1kV air variable
CR	1000 PIV/2.5A (RS 276-1114)
CR1	8.2V zener diode, Motorola 1N4738 (Newark)
C_{1T}, C_{2T}	(see table 1)
L_T	(see table 1)
L1-L5	(see table 2)
M1,M2	Triplett model 220G, 1 mA movements (152-0152), Bezels No. 13-238 M2 shunted to read 1.5A full scale
PC	4 turns No. 10 buss wire wound around 3 150 ohm 2-watt carbon resistors in parallel
Q1	2N3053 (RS 276-2030)
Q2	2N3055 (RS 276-2041)
RFC1	10 turns, No. 14 enameled wire on 1/4-inch diameter ferrite rod
RFC2	plate choke from Drake L7 amplifier
RFC3	11 turns No. 14 enameled wire, 5/8-inch diameter
RFC4	22 turns bifilar, 4-inches long on a 1/2-inch diameter ferrite rod
RFC5	70 turns No. 18 on 1/2" diameter nylon rod
RL1	(see fig. 6)
RL2	(see fig. 6)
RL3	4PDT Potter and Brumfield KHU17D11/24 VDC coil (Newark)
S1	2 pole/6 position, Centralab PA2003 (Newark 22F603)
S2	model 86 non-shorting switch, 3 pole, 6 position at 30 degree indexing (Radio Switch)
S3	3 pole/3 position 1 section, Centralab PS109 (Newark 22F777) or Centralab 2507 (Newark 22F414)
S4	momentary pushbutton 1 N.O., 1 N.C. ALCO 16SL-11 switch ALCO 6S-2 red lens
tube sockets:	EIMAC SK1900
tube chimneys:	EIMAC SK1906
tube clamp:	EIMAC SK1916

cabinet 7 × 17 × 14 inches, CTS model MCLS-71714, SPP-714 side panels, CP-1714 chassis plate

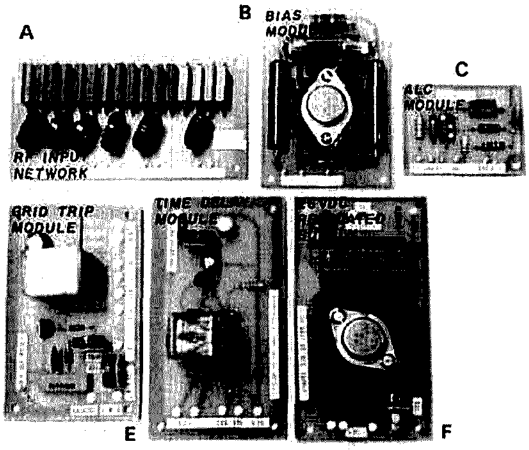
*parts suppliers.

name and address	name and address
CTS Interfab 660 Lenfest Road San Jose, California 65133 Tel: 408-251-1600	Newark Electronics 500 N. Pulaski Road Chicago, Illinois 60624 Tel: 312-638-4411
	Radio Shack Local stores
Peter Dahl Transformer Co. 4007 Fort Blvd. El Paso, Texas 79930 Tel. 915-566-5365	Triplett Corporation 286 Harmon Road Bluffton, Ohio 45817 Tel: 419-358-5015
W.W. Granger (Dayton) P.O. Box 44289 Columbus, Ohio 43204 Tel: 614-276-5231	Radio Switch Corp Marlboro, New Jersey 07746 Tel 201-462-6100
R.L. Drake Co. 540 Richard Street Miamisburg, Ohio 45342 Tel: 513-866-2421	RS = Radio Shack™

is created with the trim pot R2, which can be used to adjust the level at which the circuit trips (Q1 is turned on).

If Q1 turns on, relay RL3 actuates, grounding the relay coil and taking the load off transistor Q1. If this feature were not provided, the transistor would start gating the amplifier on and off. Therefore, it is essential to latch the grid trip relay closed. When RL3 is closed another set of contacts breaks the VOX line, disabling the amplifier. Another set of contacts applies power to the grid trip reset pilot light on the front panel indicating to the operator what has happened. Pushing the reset button on the front panel breaks the coil line, which in turn unlatches relay RL3 and puts the amplifier back into normal operation. (See grid trip module PC board and component layout.)

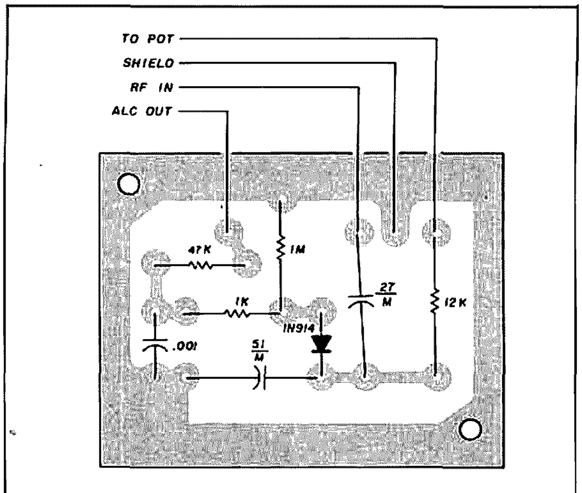
I have used this circuit in several different amplifiers. It is a good addition to any amplifier that you might build considering the high price of tubes.



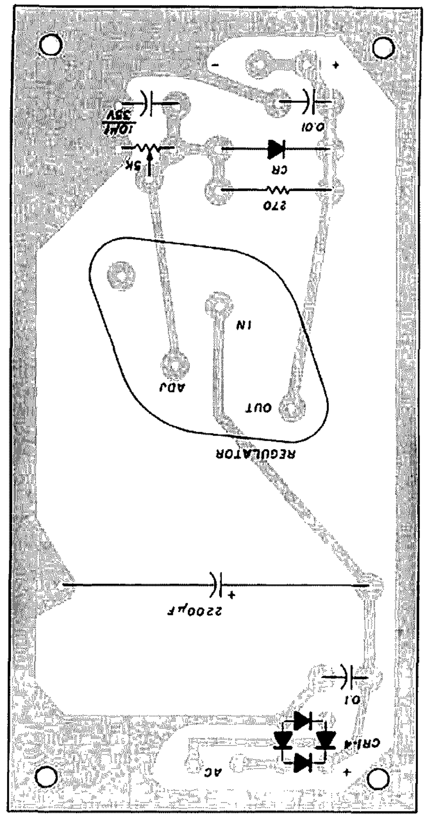
All circuit boards are made from glass epoxy boards and laid out using dry transfers from Radio Shack. (Letter codes refer to ready-to-use PC board artwork printed on page 139.)

solid-state timing module

The 3-minute timing circuit used to insure proper conditioning of the 3CX800A7s cathodes is shown in **fig. 4**. When the 26 VDC power supply comes on, current begins to flow through the 5.5M resistor R1,



ALC module (component view).

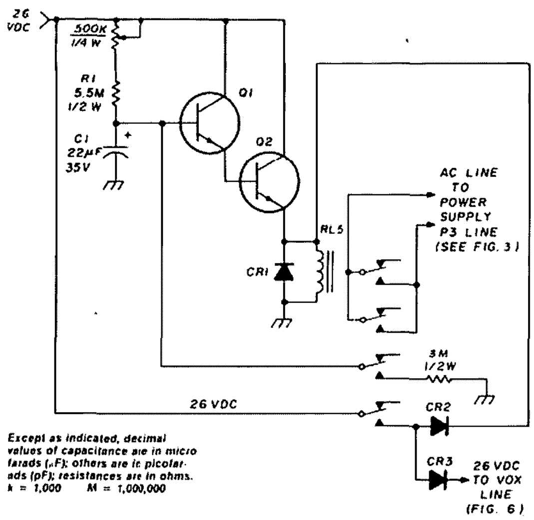


26 VDC regulated power supply (component view).

charging the 22 µF capacitor, C1. As C1 charges, the voltage is applied to the base of a Darlington configuration (Q1 and Q2) which forms an emitter follower circuit. Therefore, the emitter of the Darlington follows the voltage charge on capacitor C1. The Darlington is necessary to present a high impedance to C1 which would otherwise drain through the transistor. Relay RL5 is a 12 VDC relay that actuates at approximately 8.1 volts. Therefore when the voltage on C1 causes the emitter of Q2 to reach 8.1 volts, relay RL5 actuates allowing amplifier operation. Of course, this takes approximately 3 minutes.

Relay RL5 is a 4PDT relay. Two poles are wired in parallel and apply power to the HV power switch located on the front panel. Therefore, the HV power supply cannot be turned on during the warm-up period. Another set of contacts on RL5 is used to connect a 3-megohm resistor to C1 to discharge the capacitor thus resetting the timer module. This is a protective circuit to insure that a 3-minute warm-up cycle occurs should the amplifier be turned off for a short time and then turned back on. The fourth set of contacts serves three purposes. First, the meter lights come on, indicating an amplifier-ready condition. Second, 26 VDC is sent to the hot side of relay RL5 in the timing module itself thus latching the relay. This takes the power load off the Darlington and allows the reset function discussed above. Diode CR2 prevents the voltage on the VOX line to rise with the emitter on Q2. Third, 26 VDC is sent to the VOX line to allow the amplifier to be keyed up.

Note that it is extremely important to get a high impedance Darlington with very low leakage current to make this circuit work properly. I originally tried three different single package Darlingtons with no success. The transistors used are readily available



Except as indicated, decimal values of capacitance are in micro farads (µF); others are in picofarads (pF); resistances are in ohms.
k = 1,000 M = 1,000,000

item	description
CR1,2,3	2.5A/1000 PIV (RS* 276-1114)
Q1	2N2222A (RS* 276-2009)
Q2	2N3053 (RS* 276-2030)
RL5	4 PDT, 12 VDC coil (RS* 275-214)

*Parts supplier

fig. 4. Solid-state time delay schematic.

from Radio Shack and I recommend you use these exact components. (See timer module PC board and component layout.)

operating bias circuit

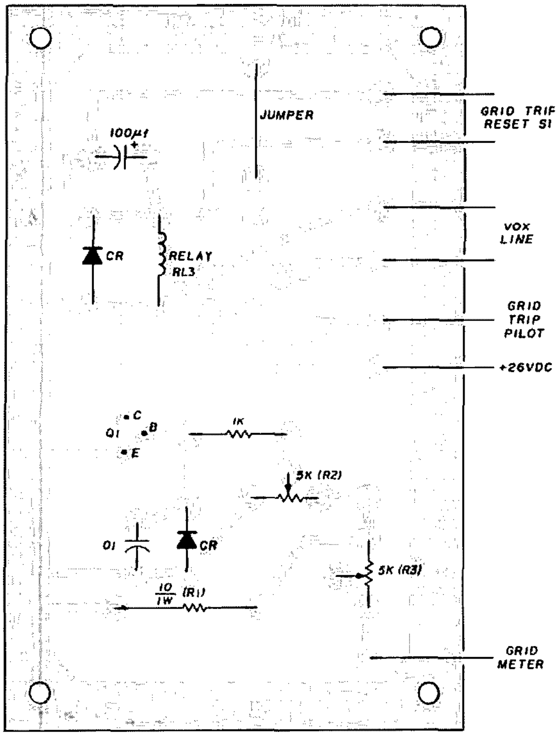
Operating bias is generated through use of a high-power 2N3055 NPN transistor (Q2) which is biased by a 1-watt zener diode to function as a high power zener (see **fig. 1**). This circuit includes readily available components and provides an easy way to adjust the bias voltage merely by changing the 1-watt zener diode between the base and collector of the 2N3055. This amplifier is biased with 8.2 volts which results in a 40 milliampere zero signal resting plate current.

This circuit is much easier to work with than the more conventional 50-watt zener diodes. Here in the Midwest the 50 watt TO3 case zeners are very expensive special order items.

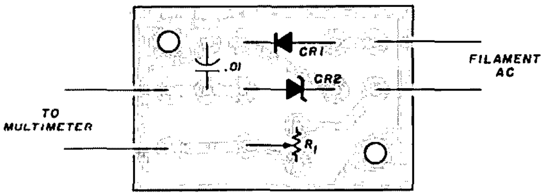
The bias circuit also has a 1.5 amp fuse to protect against excessive current. The current flowing through this circuit is the sum of the plate and grid current drawn by the tubes.

In the standby mode, the amplifier is biased to cut off (i.e. zero static plate current) by voltage generated by the current flow through the 39K resistor, R4. The resistor is shorted in the transmit mode by a set of contacts on the RF input relay RL1.

The bias circuit in the amplifier has been constructed as a single module. The module is located on the side wall of the cabinet next to the tube sockets and RF choke RFC5. (See bias module PC board and component layout.)



Grid trip module (component view).



Filament voltage meter module (component view).

metering circuits

Metering is provided for plate current, grid current, high voltage, and filament voltage.

Plate current is monitored at all times. This meter is in series with the B – lead.

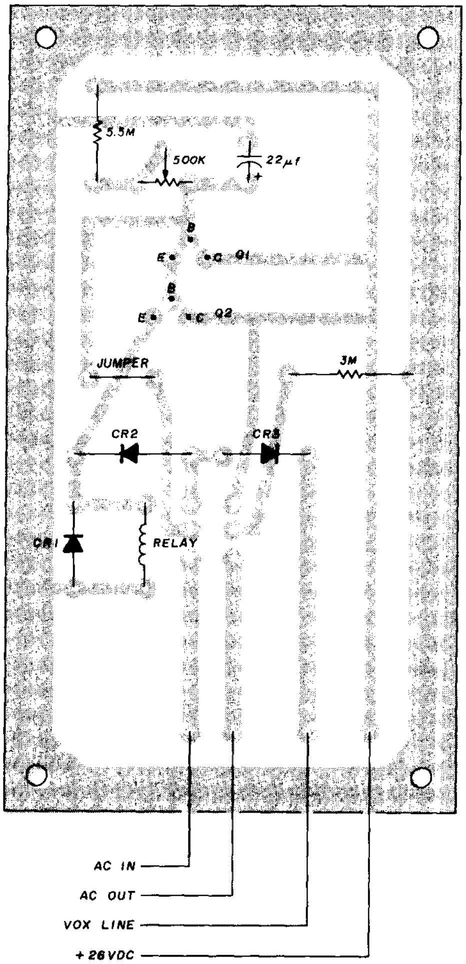
The other metered values are selected on the multimeter. The meter can be anything from a 100 microampere meter to a 5 milliampere meter full scale. You can make any of the meters work by choosing the proper calibration resistor. I used a 1 mA movement.

Grid current is measured by monitoring the voltage across the 10-ohm resistor (R1) through which grid current is drawn (refer to the discussion on the grid-trip module). Trim pot R3 is used to calibrate the meter to read full scale when 100 milliamperes is drawn through R1.

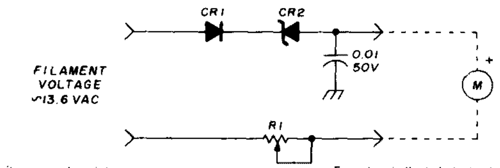
High voltage is measured by monitoring a low voltage value developed by a voltage divider in the power supply. This voltage should be kept very low since it is sent through the control cable that connects the amplifier to the power supply.

Filament voltage is very simple to monitor (see **fig. 5**) and often an ignored value in amplifier designs. A filament voltage that is too high can lead to premature tube failure. The meter has been calibrated from 10 to 15 volts. A 9.1 volt zener diode and a silicon diode are used to convert the AC filament voltage to DC and allow the bottom scale on the meter to be approximately 10 volts. No current flows in the metering circuit until the voltage across the zener diode exceeds 10 volts. (See filament voltage metering PC board and component layout.)

Labeling the meters takes a lot of patience but really contributes to the appearance of the amplifier. It is necessary to choose a meter with an analog scale that has the correct number of divisions. However, the meter labeling makes no difference. In a very clean environment, remove the meter plate from the meter.



Timer module (component view).



item	description
CR1	1 kV/2.5A (Radio Shack #276-1114)
CR2	9.1 volt Zener diode (Radio Shack #276-562)
M	1 mA multimeter
R1	25K potentiometer/1/4-watt

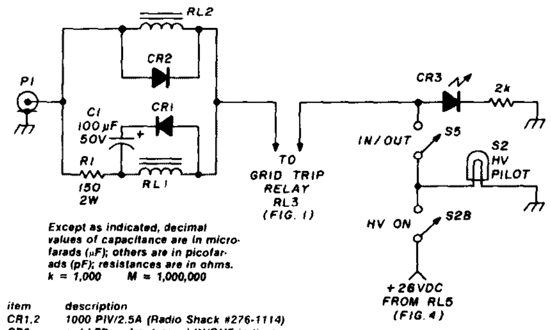
Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

fig. 5. Filament voltage metering circuit.

Any markings on the meter can be removed with a pencil eraser. Rub lightly, but be persistent. The markings will come off the face leaving a clean surface to which the number and letter markings can be applied. I use dry transfer lettering to mark the meter plates to reflect the scales I want to read. Dry transfers are now available from Radio Shack but the letters may be too large for small meters. Varied assortments of smaller letters are available from most art supply stores.

RF relay sequencing

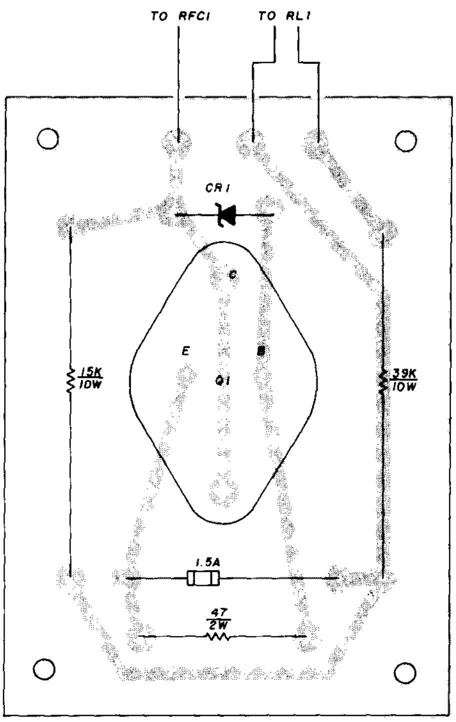
It is important to properly sequence the input RF relay (RLI) and the high power output vacuum relay (RL2) to insure that the antenna is always connected to the amplifier before RF drive power is applied. This



Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

item	description
CR1,2	1000 PIV/2.5A (Radio Shack #276-1114)
CR3	red LED — front panel IN/OUT indicator
P1	RCA plug/rear panel
RL1	4 PDT Potter & Brumfield KHU17D11/24 VDC coil (Newark)
RL2	SPDT vacuum relay, 26 VDC coil (Hamfest)
	alternate — Potter & Brumfield KUP5D15/24VDC coil (Newark)

fig. 6. RF relay sequencing circuit.



Bias module (component view).

is accomplished by closing the output relay RL2 slightly before the input relay RL1.

The R1-C1 combination, shown in **fig. 6**, causes RF relay RL1 operation to be delayed with respect to RL2.

I checked the timing by applying a small voltage across the contacts of both RL1 and RL2 and watching the voltages on the scope as the amplifier was keyed up. The output relay RL2 closes approximately 25 milliseconds before the input relay RL1. The timing constant provided by R1-C1 is subject to change depending upon the type of relays used for RL1 and RL2.

lead filtering

All control and power leads entering or leaving the RF deck are filtered. This is accomplished through the use of small coils made by winding 10 turns on a 1/4-inch (6.35 mm) diameter ferrite rod. A bypass capacitor is included on each coil to ground. Locate each filter as close to the rear panel as possible.

Also, feedthrough capacitors are used to filter all leads from the under chassis RF section of the amplifier to the front section meter compartment (see bottom view of the amplifier). All circuits that are sensitive to RF are mounted in this front compartment.

Pi-L tank circuit design

The tank circuit uses the popular Pi-L design for two reasons. First, a Pi-L gives approximately 15 dB better attenuation of the second harmonic over a more conventional Pi design. Secondly, a Pi-L allows use of a lower value plate tuning capacitor for the circuit.

The design parameters for the Pi-L circuit are provided in **table 1** (see also reference 3). The values shown in **table 1** are for a plate impedance of 1200 ohms.

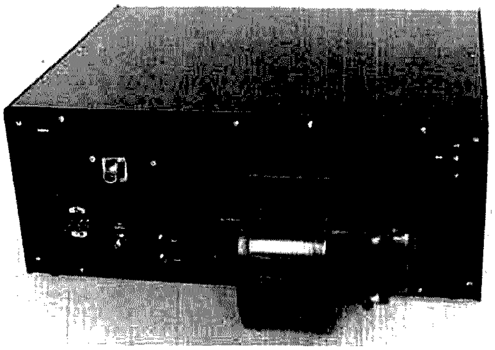
$$\textit{plate impedance} = \textit{plate voltage}/(1.57 \cdot \textit{plate current}) = 2250/(1.57 \cdot 1.2) = 1194.3$$

Because of a limitation of 500 pF for the plate tuning capacitor C1, the design for 160 meters is for a plate impedance of 2500 ohms. This translates to 570 mA plate current at 2250 volts or 1289 watts input. This provides about 700 watts output. I felt that this was sufficient power. If full power is desired on 160, some method of adding additional capacitance is needed.

Several coils are required to obtain the desired inductance. Specifications for these coils are contained in **table 1**. Note that the 160 meter and L coils are wound as toroids for compactness. Also, the self-shielding characteristics of the toroids help especially in isolating the L coil from the rest of the tank circuit.

It has become apparent to me that the one place most Amateurs feel uncomfortable is designing and installing the tank circuit in an amplifier. There always seems to be uncertainty about where to tap the coils to achieve the desired design no matter how many times you do it.

I first use a Heathkit impedance bridge to determine the approximate capacitance at various settings of the tuning knobs. I then take a high-tolerance fixed capacitor and connect it in parallel with the tank circuit at different points to determine approximate inductances of the coil set. This is done with the coil set in place in the amplifier, but with the plate and load tuning capacitors disconnected. The tubes are also re-



Rear view of amplifier shows Dayton 4C004 blower.

table 1. Pi-L tank circuit values.

F(MHz)	C1	C2	L1	L2	Q
1.8*	462	2121	22.02	8.90	13.1
3.5	469	1443	6.25	4.45	13.4
7.0	235	656	3.23	2.44	12.4
14.0	116	320	1.65	1.24	12.2
21.0	77	213	1.10	0.83	12.2
29.7	54	146	0.80	0.60	12.0

*R_L = 2500 ohms I = 570 mA or 1289 watts input (1200 ohms for all other frequencies)

Coil description:

28 MHz — 3/16 inch tubing; 3-1/2 turns with 1-3/4 inch diameter

21 to 3.5 MHz — 3/16 inch tubing; 14 turns with 2-3/4 inch diameter

taps: 21 MHz — 1-1/2 turns
14 MHz — 3-1/2 turns
7 MHz — 7-1/2 turns
3.5 MHz — 13-1/2 turns

1.8 MHz - toroid; 3 × T200-2s covered with fiberglass tape; 26 turns

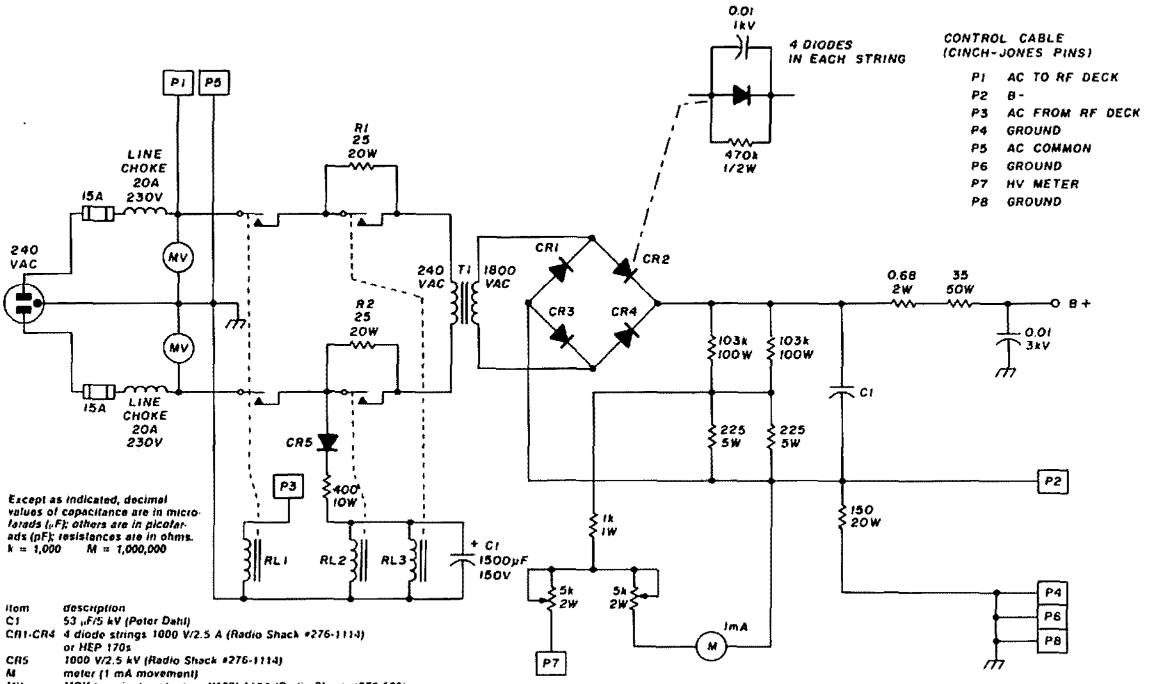
L-coil

28 to 21 MHz - No. 12 tinned wire; 1-1/8 inch diameter, 5 turns

taps: 28 MHz — 3 turns
21 MHz — 5 turns

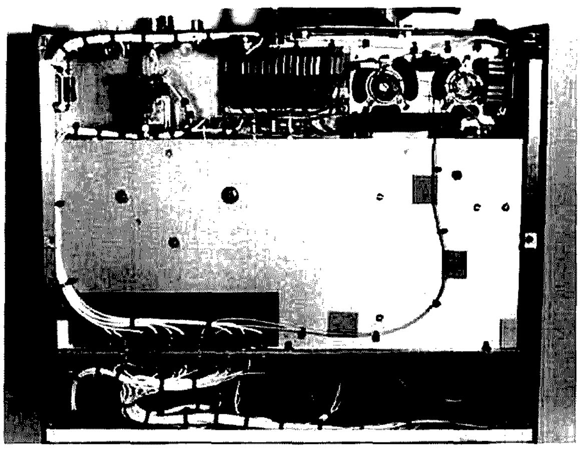
14 to 1.8 MHz — toroid: 2 × T200-2S covered with fiberglass tape; 17 turns

taps: 14 MHz — 2 turns
7 MHz — 6 turns
3.5 MHz — 9 turns
1.8 MHz — 17 turns



item	description
C1	53 μF/5 kV (Peter Dahl)
CR1-CR4	4 diode strings 1000 V/2.5 A (Radio Shack #276-1114) or HEP 170s
CR5	1000 V/2.5 kV (Radio Shack #276-1114)
M	meter (1 mA movement)
MV	MOV transient protectors V130LA10A (Radio Shack #276-570)
RL1	2 PDT mercury relay (Dayton #6X598-3)
RL2,RL3	SPST/90 VDC/25A contacts Potter & Brumfield PRD1DYO (Newark)
T1	1800 VAC/1.5A (Peter Dahl)

fig. 7. 3CX800A7 amplifier HV power supply.



Bottom view: all leads enter front compartment through 0.001/500-volt feedthrough capacitors.

moved. The frequency is determined by measuring the circuit resonance with a grid dip meter. The frequency (f) is used in the following formula to calculate the inductance:

$$L = (10^{12})/(4\pi^2 f^2 C) \qquad (1)$$

The inductances required are listed in **table 1**.

The next step is to put a 50-ohm carbon composition resistor on the output of the tank circuit to simulate an antenna load. Then set the tuning capacitors at the design values for the band being considered, allowing approximately 12 pF for tube interelectrode capacitance. Using the grid dip meter, locate the tap on the coils to obtain resonance of the circuit. The approximate setting was determined by measuring the inductance at various points with the fixed capacitor, earlier.

This method of finding where to tap the coils is effective. In operation, the setting of the tuning capacitors is almost exactly where I had predicted, using the procedure described above.

I recommend that you go through this procedure even though I have presented, in **table 1**, a design complete with taps. Variations in physical layout of the inductors and stray capacitances unique to any single amplifier could affect the exact tap settings in the Pi-L tank circuit.

input network

The input network is designed for a Q of 1. The network, ganged to the main bandswitch, provides a separate pi-network section on each band. **Table 2** summarizes the component values for each pi-section.

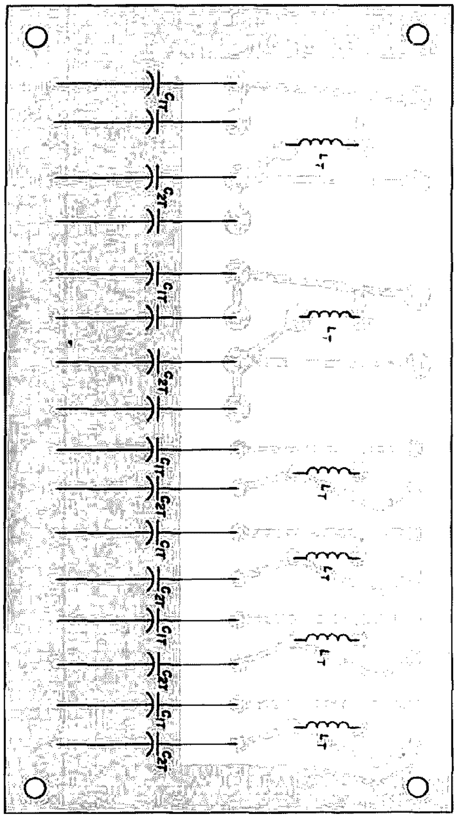
table 2. RF input network.

F(MHz)	C1*	C2*	L(μH)	No. of turns (No. 20 wire)†
1.8	1700	642	2.75	22
3.7	827	312	1.34	15
7.1	428	162	0.69	11
14.1	216	82	0.35	8
21.2	144	55	0.23	6
28.5	107	41	0.17	5

*Use standard values near these theoretical values. Anything close to table values is OK since the Q is so low.
†Coils wound on T68-2 toroid cores. $Q = 1$.

The coils are wound on T68-2 toroid cores. Input impedance for a pair of 3CX800A7s is approximately 29 ohms. The input network is important not only for impedance matching for the exciter but also to minimize distortion products.

Fine tuning of the networks can be accomplished by either spreading or compressing the coil turns on each core for minimum SWR on each band, respectively. Exact replication of the network from **table 2** may still require some small adjustment due to variations in component tolerances and differences in physical layout.



RF input (component view).

The network is built as a module on a separate PC board. A board layout has been provided.

cooling

Sufficient cooling is essential in any high power amplifier.[5] The 3CX800A7s require forced-air cooling to maintain the anodes and seals of the tubes at safe operating temperatures.

The specifications for cooling are summarized as follows:

5000 feet above sea level		
dissipation (watts)	flow rate (cfm)	water pressure (inches)
400	7	0.10
600	14	0.23
800	23	0.57

Refer to the 3CX800A7 technical sheet[4] for more information on cooling at different altitudes.

The blower used in this amplifier is a Dayton model 4C004 which is capable of supplying 45 cfm at 0.4 inch static pressure. The operating speed of the blower is 2880 RPM, which is slow enough to assure relatively quiet operation.

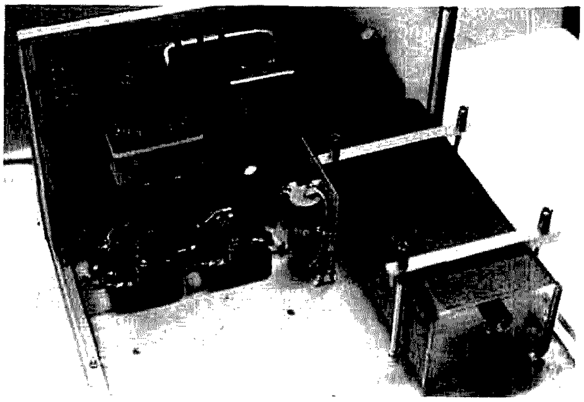
To obtain the legal 1500 watts output requires that the amplifier run at approximately 2500 watts input assuming 60 percent efficiency. This means that the amplifier is dissipating about 1000 watts which is well below the capabilities of the tubes.

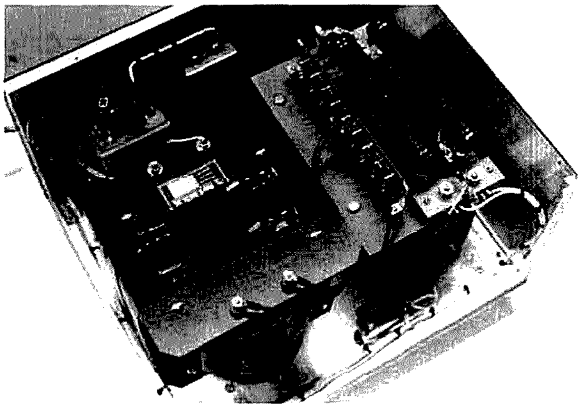
The blower mounts on the rear panel of the amplifier and blows air into a pressurized chassis. The air vents up through the tube anodes and out the top of the cabinet. I plan to eventually make a duct flange to mount in place of the blower on the rear panel and remote the blower where it can not be heard. Flexible hose will duct the air from the remote blower to the amplifier.

high-voltage power supply

One key to optimal performance of any amplifier is the power supply. It takes a supply that not only holds regulation under maximum current draw but also can supply the needed current. This translates to a quality transformer and plenty of filter capacitance. This is the area in which many commercial amplifiers fall short.

The circuit for the power supply is shown in **fig. 7**. The supply operates off of 240 VAC and incorporates a step-start circuit. When the HV switch is pushed on the front panel of the amplifier (assuming the warmup time delay has expired), 110 VAC is returned to the power supply to actuate the mercury plunger relay, RL1. This sends 240 VAC to the HV transformer through the dropping resistors R1 and R2. This avoids





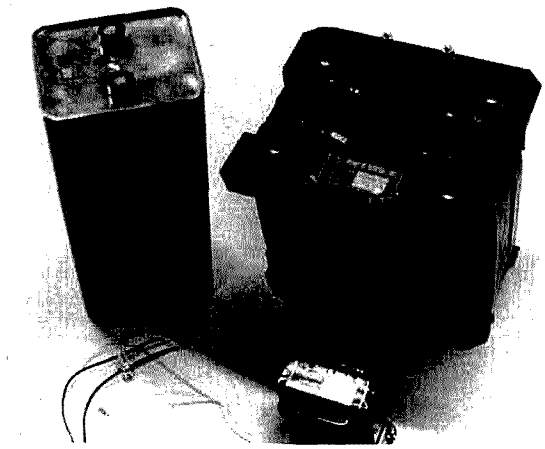
2500-volt high-voltage power supply. Cabinet is constructed from 1/8-inch aluminum panels joined with 1/2-inch angle stock.

a current surge in the supply to charge the 53 μF filter capacitor, C1. A surge could damage the diode bank. After approximately 2 seconds, relays RL2 and RL3 actuate to short out resistors R1 and R2 thus turning the HV supply on to maximum voltage.

The 2-second delay is accomplished by the time constant set up by R3 and C1. 110 VAC is rectified using diode CR5 and proceeds to charge capacitor C1. When the DC voltage on C1 reaches approximately 60 VDC, the DC relays RL2 and RL3 close (relays RL2 and RL3 have 90 VDC coils). This takes about 2 seconds.

Metering is accomplished across the voltage divider created by the 103K bleeder resistors and the pair of 225 ohm resistors. This drops the voltage in the control cable to the amplifier to approximately 5.5 volts. A voltmeter is also included in both the RF deck and power supply cabinets. Note that the meter calibration pots must be adjusted together since they affect one another.

The plate transformer used in the HV power supply was supplied by the Peter Dahl transformer company. A 2.8 kVA CCS hypersil unit is designed specifically for a pair of 3CX800A7s; the unit, now a stock item, is available at a reasonable price. Under no-load con-



Custom plate and filament transformer, 53-μF filter capacitor. (Available from Peter Dahl Co., Inc.)

ditions, the supply provides 2520 volts. Under a load of 1.2 amperes, the transformer output voltage drops to about 2300 volts! If you want the most out of an amplifier like this one, you need a power supply with a transformer of this high quality.

Peter Dahl also stocks a filament transformer for a pair of the 3CX800A7s (13.6 volts at 3.0 amperes). Both the filiment transformer and the plate transformer are well worth the investment. At the time of this writing, Peter Dahl also stocked the 53 μF filter capacitors rated at 5 kV.

concluding remarks

Building around the 3CX800A7s is a real pleasure. Their compactness and low voltage requirements open up a wide range of projects for compact amplifiers that do not have to sacrifice power capability. The thing that makes the tubes unique is that they are designed for full power operation up to at least 350 MHz.

My thanks to those who contributed so much to this project: Rich Rosen, K2RR, Bill Orr, W6SAI, Dan Redman, K8DR, and Peter Dahl. And last and most important, Jim Garland, W8ZR, who taught me how to build.

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2. Jerry Pittenger, K8RA, "A Structured Engineering Approach to the Design and Construction of Electronic Equipment," *QST*, August, 1983, page 18.
3. Irvin M. Hoff, W6FFC, "Pi Network Design," *ham radio*, June, 1978, page 52.
4. EIMAC Varian, *Technical Data Sheet for 3CX800A7, document 357655/VA4588*, 301 Industrial Way, San Carlos, California 94070.
5. Vaughn D. Martin, "Cooling Semiconductors," Parts 1 and 2, *ham radio*, July and August, 1984.

Turn to page 139 for a complete set of ready-to-use PC board artwork (foil side) for the project described in this article. — **Editor**

ham radio

junk-box ingenuity:
how to buy, use, and recycle surplus electronic parts

Don't throw away resistors and capacitors — *cook them*

At one time, a well-stocked junk-box was a ham's pride and joy. Parts scavenged from discarded televisions, radios, and record players could be used to build anything from test equipment to transceivers. Using vacuum tubes, 1/2 watt 10 and 20 percent resistors, ceramic disk and dipped mylar capacitors, almost any circuit could be built for use up through 450 MHz.

Today's junk boxes are different; chances are they contain odd parts from scavenged surplus boards or parts bought by mail or as assortments at hamfests. They contain various types of semiconductors, 5 percent tolerance resistors, electrolytic and tantalum capacitors, as well as the old stand-by ceramic disks.

Despite the differences in content, all junk boxes are alike in one regard: by definition, they are where parts are stored for long periods until used. Storage takes a toll on electronic components, although much of it is reversible; before examining this, let's consider the parts themselves.

The average ham always seems to be looking to buy new, prime, MIL-spec parts at hamfest prices. Unfortunately such parts are usually not available to the small buyer at any price, although some manufacturers will sell directly to individuals. As a result we buy blister-packed parts at Radio Shack, or by mail from any number of suppliers. While nearly all of these dealers are reputable, the parts you get may not always be exactly what you originally had in mind.

what is "surplus"?

Consider, for example, the origin of parts termed "surplus." Sometimes a company accidentally over-purchases some new parts and instead of paying a re-stocking charge, elects to sell them to a surplus agent; *these are the parts you want.* Sometimes a manufacturer discovers that a part from one company — often a semiconductor — works in their circuit, while another company's equivalent doesn't. If no other use for them can be found, they will either be thrown away or sold as surplus. (Since the fault lies within the idiosyncracy of the design, the parts are usually not returnable.) *These may also be good parts.* An important point to remember is that most large buyers of parts have sufficiently amicable relations with their suppliers to return *good* parts for credit.

There's another end of this spectrum: from time to time a manufacturer buys parts that don't meet specifications, but can't be returned for some reason. These will *occasionally* end up on the surplus market. A component manufacturer, for example, may produce a run of products that is not in "spec." While these are most frequently discarded (although garbage cans are sometimes raided) they may be sold to the surplus market. It has even happened that manufacturing runs of semiconductors have been stolen before testing and culling of rejects can be completed; obviously, you don't want *these* parts.

The surplus boards on the market are of a few basic types:

- those that have lived out their expected life and are now experiencing a high failure rate;
- boards made obsolete by a change in design;
- seriously damaged and therefore non-repairable boards (damage can include degradation of critical areas, such as gold fingers; submersion in a solder wave; cuts from a mass lead trimmer; or other actions rendering the board electrically unfixable);
- boards simply overbuilt by a company; now and then a manufacturer will sell boards from inventory to raise some cash.

While surplus boards should not be expected to perform their original design functions, chances are that most of the components on all but the first type listed are usable. Even the unfixable boards rarely have more than a few bad components.

By Bob Lombardi, WB4EHS, 1874 Palmer Drive, Melbourne, Florida 32935

In general, then, there's no telling where "grab-bag" parts come from. They might be old or new, merely extras, or genuine junk. Given the diversity of parts from so many varied sources, it's no wonder that we occasionally build a project and find that it's a bad component that keeps it from working.

Take a clue from industry, then, and test your parts before using them. For ICs this may be impractical, but for many types of transistors, diodes, and, of course, resistors, capacitors, and coils, it is entirely reasonable.

are they out of spec?

Chances are that if you took out a DVM and tested every one of your ±5 percent carbon composition resistors, you'd find plenty of them out of tolerance. In particular, they'd all be on the high side of their marked value, perhaps as much as 7 to 10 percent above it. *Don't* throw them out.

The reason for these high values is something I alluded to earlier: improper storage. Carbon composition resistors are essentially carbon and impurities mixed into a binder and molded into shape. The resultant mixture is hygroscopic — that is, it absorbs moisture, which raises the resistance value. This phenomenon is entirely reversible, if the parts have not been over a flow solder (or through any other high temperature operation) while loaded with moisture.

The importance of this consideration depends on the intended use. With the exception of precise RC timing circuits, digital circuits are quite forgiving of resistor values. Analog circuits can be fussier, especially those (such as audio filters) requiring balanced parts or matched values. Most other circuits are less particular. In designing homebrew circuits, you frequently make up for these parts by various means, making the actual value, even if it is "wrong," the design value!

If you need to reverse the problem — i.e., *lower* the values — heat the parts to about 200 degrees F (93.3 degrees C) for a period of time appropriate to the size of the part: 25 hours for 1/8 watt, 50 hours for 1/4 watt, 75 hours for 1/2 watt, 130 hours for 1 watt and larger. If running your oven at 200 degrees for 50 hours or more is inconvenient or too expensive, you can construct a plywood box, line it with fireproof insulation, and add a light bulb or two for temperature control. Some experimenting is necessary for a controlled temperature, but I've seen similar boxes used by small companies for many purposes.

If you can't control the temperature accurately, err on the side of too cool rather than too hot; waterlogged parts heated beyond the boiling point of water will explode like popcorn. Using lower temperatures will necessitate leaving the parts in longer; just how long can be determined by removing parts from time to time and checking values.

handling capacitors

Manufacturers of ceramic capacitors also recommend a heat soak of a day or two at 200 degrees F prior to critical value measurement, although I've never seen a problem with this in industry. One manufacturer recommends heat soaking at 125 degrees C for 4 hours or 150 degrees C for 1/2 hour. This "de-ages" the part to its original value and effectively begins its life anew.

Electrolytic capacitors deform with time as a result of a breakdown of the dielectric layer, and should be reformed before use. This is done by applying the rated voltage to the part for 30 minutes through a current-limiting resistor; 1 K ohm is adequate for up to 100 VDC capacitors. Let the part "rest" for a day at room temperature.

The small aluminum electrolytic capacitors popular today may not be much of a bargain if garnered from surplus boards. There are two reasons for this: first, they have a rated life of only 3 to 5 years, if run constantly; second, cleaning these capacitors in hot vapor degreasers will cause premature failure if the parts are not sealed with epoxy over the rubber seal plug on the cap. Of course, because you never know how used parts were handled, you'll have to decide whether to use them or not.

Tantalum, silver mica, polyfilm, and most other capacitors tend to be more stable in storage if their hermeticity is good. In general, if a capacitor is so damaged that the plates are visible, it should be thrown out. Minor chips and cracks may not matter; obviously no electrolytic should be used if its case is punctured or badly dented.

active devices require special handling

Transistors, diodes, and ICs in epoxy packages are quite rugged, but only if the seals are good. Beware of ceramic or glass packages, which can have hairline cracks.

Enough has been written about static damage that all hams should be aware of it. But less widely known is electrical overstress (EOS) due to a static field, rather than a discharge. This causes operation to degrade gradually, resulting in "flaky" operation (subtle timing errors that occur intermittently, some "soft" RAM failures, and so on) until the system degrades to failure. The only solution to this problem is prevention; with parts bought from anyone other than the manufacturer, there can be no guarantee that electrical overstress has not occurred.

By the way, don't think that this problem is only found in CMOS or MOSFETs. All semiconductor families, including bipolar, have been shown to be degradable by low-level fields of less than 1000 volts.

(Actually, once you've read some of the literature available on this topic, you begin to not want to handle semiconductors unless you've strapped yourself to a conductive table top with No. 2 battery cables.)

If you've measured parts and found them to be well out of spec, it may pay to consider the accuracy of your measuring device. Your meter (DVM, cap meter, LCR bridge, or other instrument) also has accuracy tolerances, and they can overlap with the part being measured. If your meter is rated ±3 percent, for example, and is running at the high end of its tolerance, any parts that are only 2 percent higher than their rated value will appear to be *5 percent* above, and thereby possibly out of spec. It's no wonder that calibration is so highly stressed in high quality operations!

Yet another factor is test method. Consider a common case: that hypothetical DVM you're using probably applies about 1 to 1.5 volts to a resistor during ohm measurement, and reads the current flowing in the circuit. With low value parts, such as those of 10K or less, this is fine; assuming 1.5 volts, the current being measured is 150 microamperes. With parts over 100K, say 1 megohm, the current is 1.5 microampere; this is somewhat harder to measure accurately, especially the detection of small differences in current. This is one reason why exacting specifications for resistor tests call out higher voltages for measuring larger value resistors.

Although this article might seem to emphasize problems — first telling you to measure your parts, then urging you to doubt your results — that's not at all my point; what I'm suggesting is that in order to buy parts intelligently, it helps to know where they came from and what to do with them when problems arise.

It seems strange to me that homebrewing should be on the decline today; this should be a "Golden Age" of homebrewing if there ever was one! Why? Just look at what's available! ICs that perform all manner of digital and analog functions are at our disposal, and most are quite cheap. Using reasonable parts counts, we can build circuits capable of performance that was no more than the stuff of dreams in the 1950s and 1960s; in the 50s, who would have dreamed of a 3-terminal voltage regulator! We can even build things more cheaply today than we could then; just look at old issues of the several Amateur Radio magazines, and you'll find that the *dollar* prices of gear are essentially the same now as they were in the mid-1960s; taking inflation into account, prices have come *down* considerably, while performance has gone *up*.

Keep a well-stocked junk-box, know what's in it, and do your part in restoring the Golden Age of homebrewed gear!

ham radio

constructing air-wound coils

Build your own for high Q, low cost

For the lack of a proper coil, a project such as an amplifier, tuner, or antenna, may be slowed down or stopped. If your local supply house doesn't stock coils and if you don't want to order by mail, you can wind your own.

Using this simple technique, professional looking air inductances (**fig.** 1) 2 to 4 inches (5 to 10 cm) in diameter with wire sizes No. 8 to No. 16, can be fabricated, at a cost of no more than about 25 percent of commercial coils.

easily obtained materials

This method is based on the use of a wood cylinder commonly called a *mandrel* (**fig. 2**). **Fig.** 3 shows the dimensions of wood blocks which will produce a mandrel to fabricate 3-inch (7.6 cm) diameter coils. I have three such mandrels, 2, 2-1/2, and 3 inches (5, 6.4, and 7.6 cm) in diameter. This selection accommodates all my coil needs. **Fig. 4** shows the dimensions of a 3-inch (7.6 cm) diameter mandrel, in finished form.

Several choices of wire are appropriate. You can use the bare, soft copper wire (grounding wire) available in many sizes at building supply houses, or plastic covered solid wire stripped of its insulation. Enameled wire works well, but may present a problem if you intend to solder taps on the coil; the enameled insulation is difficult to scrape off. Tinned copper makes especially good looking coils, but while it's easy to solder, it may be difficult to obtain.

The coil supports are glass epoxy board cut into strips 12 inches (30 cm) long, 1/2-inch wide (13 mm) by 1/16-inch (1.6 mm) thick. I use discarded printed circuit board material, which can be cut easily with a hack saw. Just insert the blade of a knife under the copper foil and peel it off. Any remaining holes in the board will be filled with epoxy cement.

Sand one side of the PC board strips to remove the glaze and improve the bond between the cement and board. Do not use plexiglass or equivalent, which will deteriorate with exposure to heat or sunlight.

Other materials needed are 5-minute epoxy cement, kitchen type wax paper, some wood screws, a few rubber bands, and twine or plastic covered hook-up wire, the diameter of which will determine the wire spacing.

fabricating the mandrel

Fig. 3 shows two blocks of wood held together with wood screws. The screw heads are counterbored deeply enough to prevent interference with the turning operation on the lathe. Place the assembled block in a metal or wood turning lathe on the indicated centers. Shape this assembly into a tapered cylinder — while the tapered dimensions are not critical, tapering *is* necessary — per **fig. 4**. This is a simple job. If you don't have a lathe, a friend with a lathe should be willing to do this for you.

Grooves are cut using a table saw with a dado cutter of correct width. The insulating strips must fit loosely in these grooves; observe the tolerances shown in **fig. 4**. Grooves can also be cut with an ordinary blade on the circular saw by setting the blade to the proper depth. Cut one blade width at a time, rotating the mandrel to widen the groove progressively until it is wide enough. Because of the limited space available on its circumference, a 2-inch (5 cm) diameter mandrel can have only three grooves. These cannot cross the diagonal split; if they did, the mandrel would not slide apart. The finished mandrel is a cylinder split diagonally and tapered from one end to the other. It will slide apart freely after the coil has been wound and cemented.

winding the coil

Take the mandrel apart and insert three layers of wax paper between the two halves. This paper will allow the mandrel to slide apart easily. Put the man-

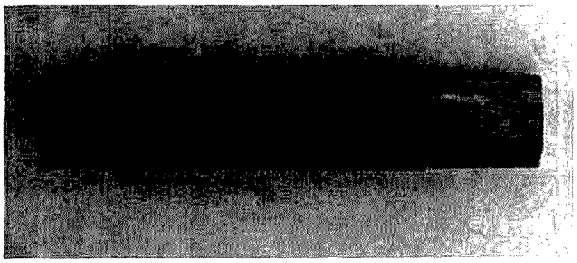

fig. 1. Finished coil on mandrel.

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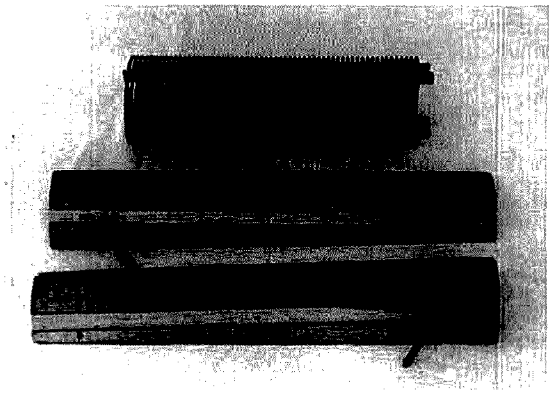

fig. 2. Finished coil with the mandrel removed and disassembled.

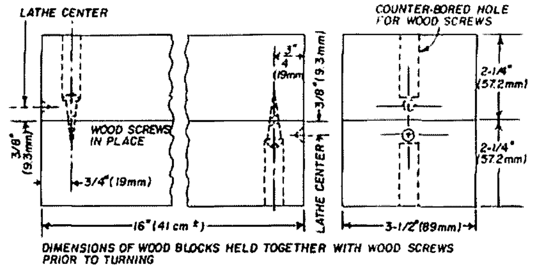

fig. 3. Wood blocks before turning. Offset turning centers are necessary to obtain a tapered split.

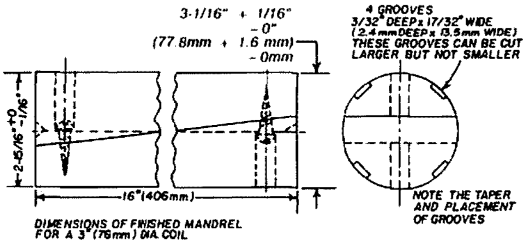

fig. 4. After the lathe turning operation, the mandrel is round and slightly tapered.

drel together with the screws and cover it with wax paper, allowing an overlap of about 1.5 inches (3.8 cm). Lay the glass epoxy strips (or cut epoxy board strips) in each groove, sanded side up, over the wax paper; they should fit in the grooves loosely, with play in the width as well as depth. Slip several tight rubber bands over the mandrel to keep the wax paper and strips in place. Center the strips and paper on the mandrel *lengthwise*.

Calculate the length of wire and twine needed to wind a 12-inch (30 cm) long coil by multiplying the number of turns you need times circumference of the mandrel. While you may not need this much coil, the leftover stock can be used for your next project and it's better to cut a piece too long than too short. Anchor one end of the wire securely, working outdoors if necessary for sufficient work space. Stretch the wire to remove all bends and kinks, then polish the wire with sandpaper for an attractive appearance. The epoxy cement will adhere better to the polished wire than to unpolished wire, and if you have to tap the coil, soldering will be easier.

Bend a hook in the free end of the stretched wire and fasten this hook to the wood mandrel, about 1.5 inches (3.8 cm) from one end with a wood screw. Lay out a length of twine the same length as the copper wire. Then wind the wire and twine on the mandrel tightly, parallel, removing the rubber bands as the winding progresses. The wound wire will hold the wax paper and strips in place. Wind the wire and twine tightly against each other to insure accurate spacing.

After the wire and twine are completely wound, fasten the wire to the mandrel with another wood screw, and remove the twine. Mix and apply only enough 2-part epoxy cement to cover one insulating strip sparingly, with no excess cement on the wax paper. If you've maintained the proper clearances on the wood mandrel, the insulating strip should move freely under the wire. Slide the strip back and forth, distributing the cement evenly between the wire and strip. Let it set for 5 minutes. Cement the other strips at 5-minute intervals. When cementing is complete, set the project aside overnight to allow the cement to harden completely.

removing the mandrel

Remove all wood screws anchoring the wire and holding the mandrel together. Place the mandrel vertically, larger end down, on the edge of a table. (The table should support the mandrel only up to the edge of the split.) Hit the smaller diameter split end with a block of wood. The mandrel will then slide apart easily. You now have a finished coil.

the end product

If you feel the finished coil needs more support, another layer of epoxy cement can be applied to each strip. I cement another strip on the coil for mounting purposes; this also makes a structurally stronger coil. In addition to strength and stability, this home-built coil offers high Q dissipating very little power even at the kilowatt power level. I have used coils built by this method in my trapped dipoles, antenna tuner, and final amplifier, with great success.

ham radio

improving amplifier ALC circuits: part 1

Grid current derived ALC helps upgrade AB_2 amplifier performance

Part One of this article examines current ALC techniques and illustrates methods of improvement. Part Two details several modifications (including ALC) for input matching and tube protection in the compact MLA-2500 amplifier. **Editor**

Although modern exciters contain adequate ALC circuits, modern AB_2 amplifiers lack truly automatic practical circuits. In this article, several approaches to amplifier-developed ALC are examined and a practical circuit is developed, using a grid current derived sample. This circuit is used in a Dentron MLA-2500 to protect the 8875 tubes from overdrive and grid destruction. Adaptability to other tube types is also discussed.

One of the most important assets of a modern exciter is an ALC circuit that prevents overdrive of the exciter's final and intermediate stage amplifiers. Controlling the drive — or load — helps preserve spectrum space as well as the exciter's output devices.

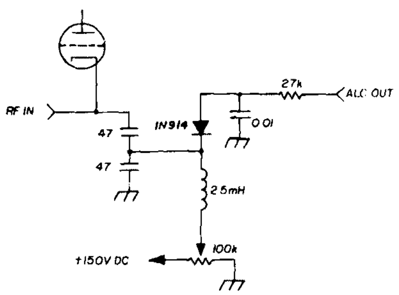

fig. 1. Typical HF amplifier ALC circuit.

Few, if any, modern exciters lack ALC. Forward power, reflected power, final amplifier current, frequency and other parameters are used to set a power output level at which the exciter can operate without distortion or destruction. It is also true that few — if any — modern linear amplifiers have a satisfactory ALC circuit to prevent overdrive from occurring. This is so even though it is as important to control drive to an external amplifier as it is to control the drive to an exciter's final stage.

The term ALC is derived from early Collins nomenclature: *A*utomatic *L*oad *C*ontrol. Many amplifiers include a circuit such as that shown in **fig. 1**. This circuit is clearly not automatic, but is based solely on RF input voltage. Such circuits offer protection from overdrive only when adjusted for each set of operating conditions on each band; they also offer no amplifier tube protection. Two important exceptions to the rule are amplifiers that do protect themselves from overdrive and consequent destruction when connected to the exciter's ALC: the Collins 30S-1 and the E.T.O. Alpha 77.

30S-1 ALC circuit

An example of true ALC is the circuit used in the Collins 30S-1 amplifier (**fig. 2**). The 30S-1 is an AB_1 amplifier; any grid current automatically indicates an overdrive condition. A 10K:10K transformer provides DC isolation between the ALC and grid circuits. In SSB service, an AC voltage is developed across the primary and secondary of the ALC transformer proportional to the grid current that flows at the audio frequency (rate) of the incoming signal. A substantial amount of control voltage is available for small values of grid current by using a voltage doubler circuit to rectify the AC voltage present across the secondary of the transformer. Sensing grid current variations is an effective way of preventing overdrive and distor-

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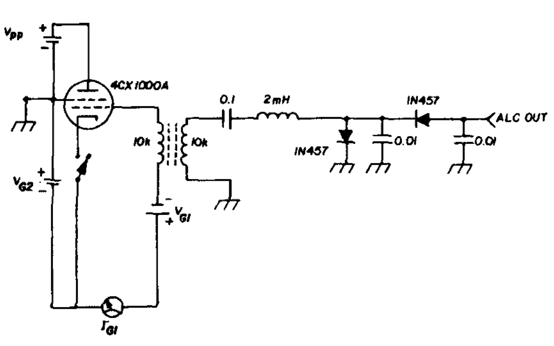

fig. 2. AB_1 ALC circuit used in the Collins 30S-1.

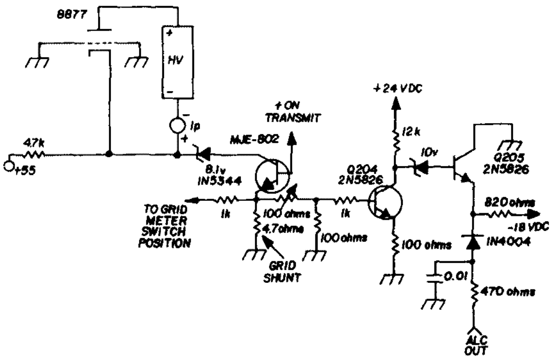

fig. 3. AB_2 grid-current-derived ALC circuit used in the E.T.O. Alpha 77 amplifier.

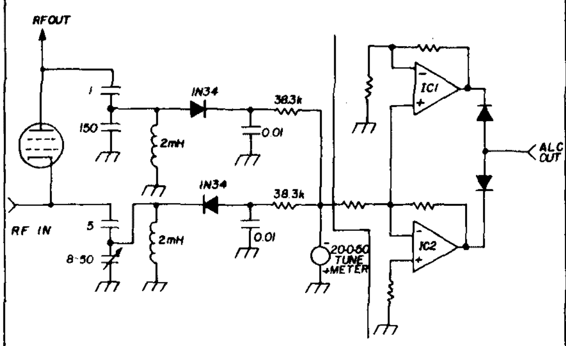

fig. 4. Comparator-type ALC circuit derived from the Collins 30S-1 tuning and loading comparator circuit.

tion in this amplifier. The particular sensitivity of the circuit is achieved by the 10 kilohm impedance of the grid transformer. This high impedance in series with the grid can cause problems if the ALC voltage is not returned to the ALC buss in the exciter. Anyone who has ever heard a 30S-1 being used without the ALC interconnected can testify to the extraordinary bandwidth that results from even minimum overdrive when the grid voltage is subject to dynamic instability.

The 30S-1 circuit is not adaptable to most modern amplifiers in Amateur use. The most common circuits use zero-bias tubes operated in Class AB_2. Significant values of grid current are necessary and normal. But there is an important analogy: just as the onset of grid current signals an overdrive condition in an AB_1 amplifier, certain fixed amounts of grid current can signal overdrive or dangerous drive levels in an AB_2 amplifier. It is quite possible to destroy a grid by applying drive without plate voltage being present. As the price of tubes soars, protective circuitry based on grid current seems a necessary adjunct for tube preservation alone even without consideration of performance.

E.T.O.* uses this type of circuit. The Alpha 77 is a modern AB_2 amplifier that has an ALC circuit designed to positively limit the 8877 grid current to 150-200 milliamperes (see **fig. 3**). Grid current is sensed, amplified, and inverted by the Q204, Q205 circuitry. In operation the negative ALC output voltage serves to limit the grid current to a preset, nondestructive limit even under conditions of mistuning, or worse, no plate voltage. In normal operation 150 milliamperes of grid current represents the upper limit of the tube's linear range.

modifying an older ALC circuit

The MLA-2500 amplifier ALC circuit shown in **fig. 1** is representative of the majority of ALC circuits used in other amplifiers. The lack of protection for the grids of the 8875 tubes offers an opportunity to adapt a circuit that protects against overdrive and potential tube destruction. An added incentive is the fact that it costs as much to replace the final tubes in an MLA-2500 as to buy a used amplifier.

The first circuit I experimented with was based on the tuning and loading circuit of the 30S-1. **Fig. 4** shows the comparator circuit which samples both input and output RF voltages to detect nonlinearity resulting from *any* cause. The use of DC amplifiers to generate a negative going ALC voltage when the comparator output voltage departs from its null seemed initially to be a satisfactory solution. In practice it failed. The impedance is fairly high and stray capacitance and inductance effects were different on each band. In addition, the circuit required correction to provide protection when the mode of operation was changed from SSB to CW which increased circuit complexity substantially. Disabling the circuit was also necessary in order to initially tune up. I soon abandoned this approach.

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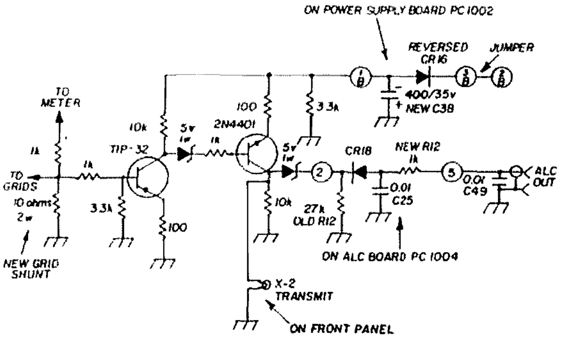

fig. 5. Final circuit as installed in Dentron MLA-2500.

a working MLA-2500 ALC circuit

Fig. 5 shows the final circuit that was developed for the MLA-2500. It operates solely through the sensing and control of grid current. I had observed nonlinearity in several different sets of tubes at more than 55-60 mA of grid current. After reviewing the tube specification sheets it soon became apparent that I could prevent overdrive and tube destruction by simply limiting exciter drive to 60 mA grid current under all conditions. The key to the circuit was a simple change in the value of the grid shunt. The new value was chosen to develop 0.6 VDC at 60 mA grid current, i.e., 10 ohms. A series multiplier resistor had to be added to the metering line. By using a 1-kilohm meter multiplier, full scale grid current was changed from 1 ampere to 100 mA. I modified the existing MLA-2500 ALC and power supply boards taking advantage of existing parts and wiring. **Fig. 6** shows the construction technique used for the remaining components added. **Fig. 7** shows the actual installation in the MLA-2500. The open area beneath the meter/function switches offered enough space to mount the component strips. After installing the ALC circuit and conecting it to the exciter, the grid current could not be driven past 60 mA. Combined with the visual warning of the front panel-mounted transmit lamp, X-2, and the increased sensitivity of the grid current meter the circuit offers positive protection. Now, if the supply voltage is accidentally left in the CW position I no longer see the high grid current that previously occurred. It is also now practically impossible to overdrive the amplifier in the SSB mode. Loading the amplifier too lightly results in excess grid current which reduces the drive and prevents flat-topping; the load is automatically controlled. It is truly an ALC circuit.

adapting to other amplifiers

Other tube circuits operating in AB_2 can also benefit from the addition of this circuit. 3-500, 813, and 811 users can all probably point to some finite value of grid current in their amplifiers which represents an overdrive condition: for example, my 3CV1500A7 (3CX1000A) begins to distort above 325 mA. The circuit is adapted to other amplifiers by simply choosing a value of grid shunt so that the desired maximum grid current develops 0.6 VDC. The circuit I developed is useful where the grid voltage sample is negative; the E.T.O. circuit could be used where the sample is positive. In either case the addition of a grid-current-derived ALC circuit represents the addition of an extremely useful operating adjunct. By using multiple isolation transistors, as shown in **fig. 8**, other parameters can be used to limit exciter drive. I have not found such circuits necessary with the 8875s, however. (A more extensive set of construction notes

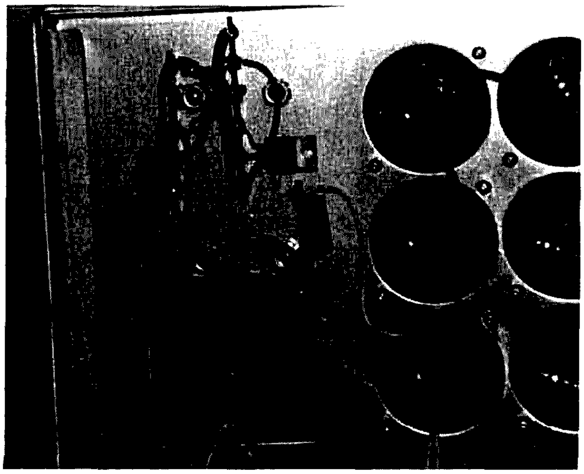

fig. 6. Terminal strip construction technique used for grid-current-derived ALC circuit in MLA-2500.

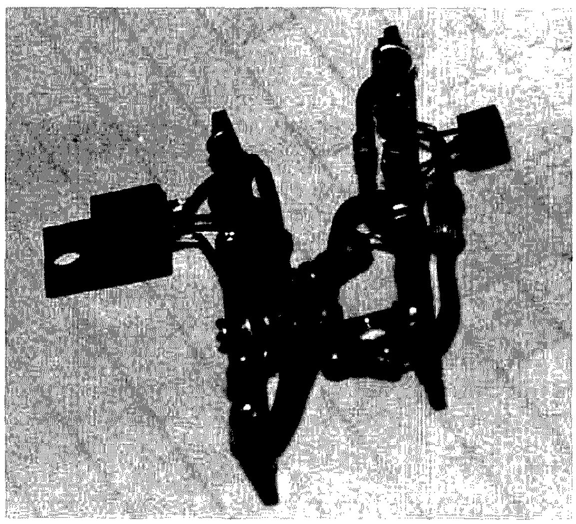

fig. 7. Final installation of ALC circuit in MLA-2500.

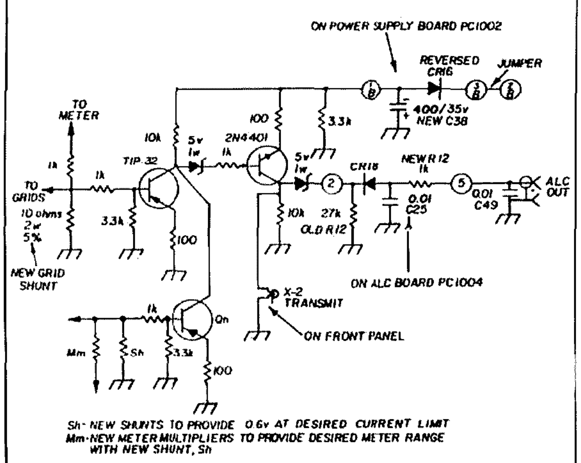

fig. 8. The ALC circuit can also be used to protect other circuit elements, e.g., plate current.

and an input matching circuit follows in Part Two of this article. — **Editor.**)

acknowledgement

I wish to thank Tom Kneadle, W8EII, for the photography, and Rodger Miller, KC8DA, for the use of his amplifier.

ham radio

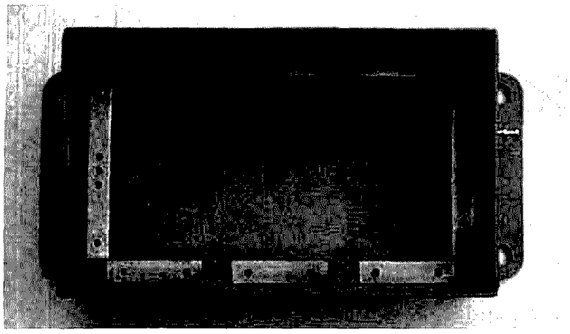

fig. 1. Interior view of box built for a wideband RF amplifier. (See *ham radio*, March 1984, pages 12-28.)

build a better box

How to modify pre-built enclosures for RF-tight design

Sometimes it seems that packaging a project is more difficult than building the actual project itself. In packaging a complete project, building the right box is critical, because if the packaging is not done well, the device inside may not work properly.

Commercially available die-cast boxes – you know, the ones that seem to be just about perfect for many things, but are never quite right for anything – appear, at first glance, to be ideal for small RF enclosures. There is no convenient way to mount *anything* in these. If you try to mount your neat, low noise, wideband amplifier in one of these boxes with a few standoffs to the bottom, you quickly find that your mounted amplifier has become an oscillator and now generates a considerable amount of output with no input.

If pre-manufactured boxes are to be used successfully for RF enclosures, particularly above several hundred megahertz, they must be modified first, with a solid, well defined grounding configuration to tightly couple the box to the circuit ground. The mechanical assembly described and illustrated in this article, *though* not particularly simple, provides an excellent packaging configuration for small RF projects through L-band. Its fabrication should be well within the capability of most Amateurs (no machine shop is required) and result in a much more successful and professional-looking project than any collection of coffee cans, PC board pieces and shim stock, however skillfully assembled.

Fig. 1 shows what your finished project should look like using this method.

drilling patterns

I used a Pomona 2901 diecast box for this article, but similar products can be adapted with equal success. The overall assembly is shown in **fig. 2**. From that you can see the relative mounting positions of the eight mounting blocks and RF connectors. The details of the box modifications are shown in **fig. 3**. All holes must be positioned with reasonable accuracy. The best way to accomplish this manually is with a precision steel rule calibrated in tenths or hundredths of an inch. This should be available from almost any hardware store or machine tool supply store. Because the diecast boxes vary somewhat in size, the holes in **fig. 3** are dimensioned from the side center lines. Experience has shown that dimensioning from one end can lead to problems with fitting parts. Before drilling, be sure to centerpunch all holes to prevent the drill from "walking."

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A number of different filters may be used. If you use other than the one listed, be sure to drill and tap the filter mounting hole to fit your particular filter.

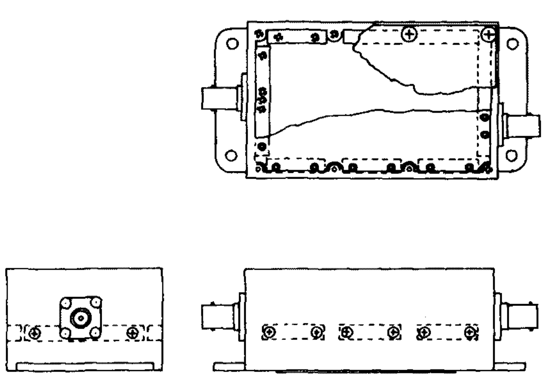

fig. 2. Assembly drawing of RF enclosure.

Don't try to use solder-in filters; the box alloy is very difficult to solder. The various screw-mounted filters are your best choice. You'll also notice that one RF connector is centered and the other offset in the corresponding ends. This is a convenient layout, but be creative and mount the connectors where best suited to your project.

treating the surface

The paint must be removed from the inside of the box where the mounting blocks are to be installed. The paint in these boxes is tenacious; the easiest method of removal is with a small sandblaster. But if you don't have a sandblaster in your work room the next best thing is sandpaper. A little work with about 120 grit wet-or-dry should do the trick. Use plenty of water to keep the paper from loading. (If you use regular sandpaper, don't use water — you'll end up with a handful of wet sand and soggy paper.)

An unpainted box such as the Pomona 2906 can also be used, but then you'll have to prepare the sur-

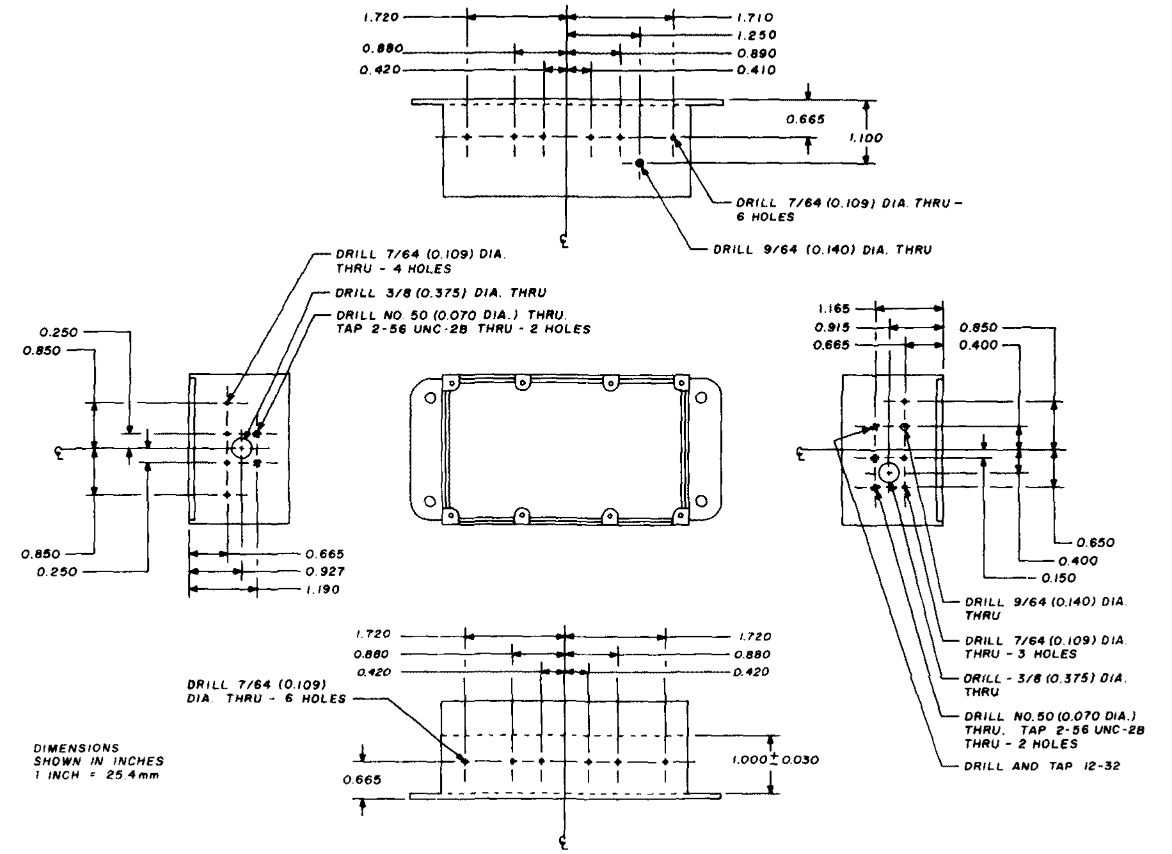

fig. 3. RF enclosure drilling pattern.

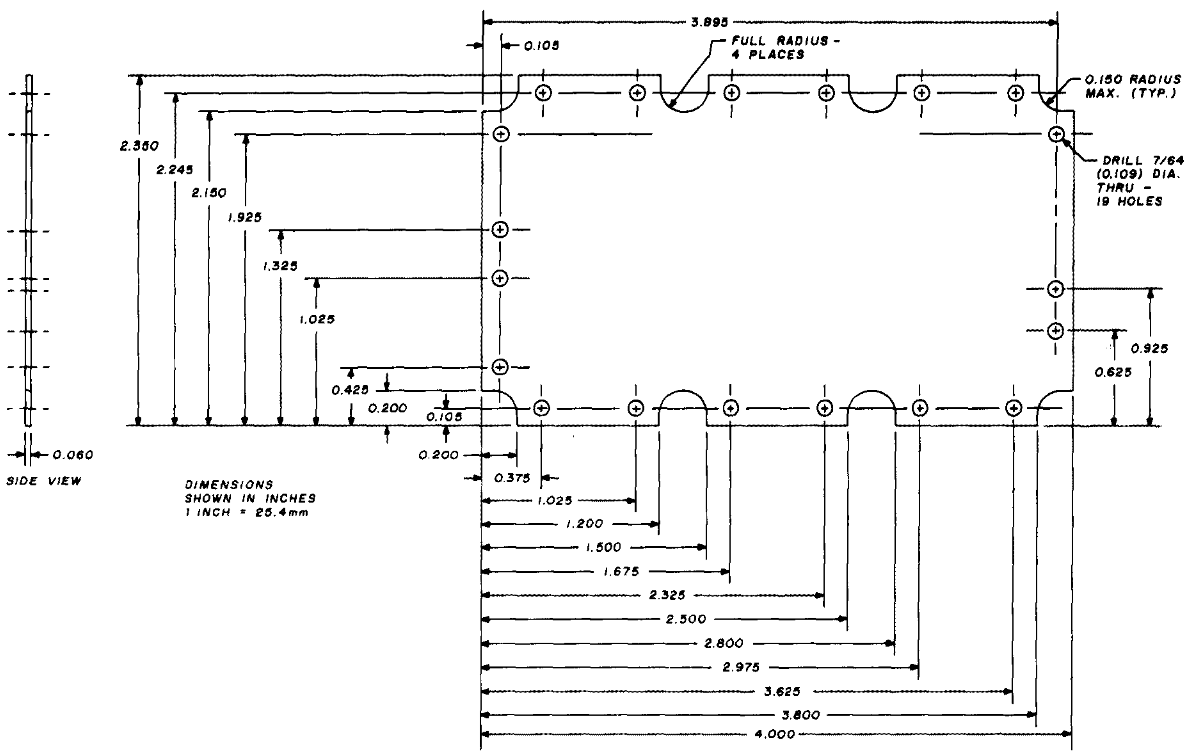

fig. 4. Board outline drawing indicates hole and cutout patterns.

face and paint the finished unit for an attractive "professional" look. This can be tricky; the bare cast box requires considerable effort if you're determined to produce a visually appealing product. All things considered, I think the painted box — with interior paint removed — is the easier to use.

interior details

Once you have all the holes drilled in the box and the paint is removed, you have to make all those little pieces that fit inside the box behind the holes. This is a tedious task, but with a little care and patience, good results can be expected. The details of the six side blocks and the two end blocks are shown in **fig. 5**. The 2-56 tapped holes are quite small, so you must be very careful when drilling and tapping — particularly in tapping. Use plenty of oil for both drilling and tapping, and be sure to back the tap out often to prevent jamming.

component assembly

After all the interior pieces are completed, the box is ready for assembly. For a little more professional appearance, use flat washers with all the screws, but be sure to mount the filter *without* a washer. On some filters, the thread is long enough to accept a nut on the inside of the box to lock the filter in place. You may have a problem purchasing the ground terminal specified in small quantities; if a suitable terminal cannot be found, a 3/4-inch, 4-40 screw and nut may be used. Place the nut about half way up the screw, drive the screw into the ground terminal location about 1/4 inch, and tighten the nut. Use a plated brass or steel screw to make soldering easier.

The finished box is now ready to have something put in it. What you install is up to you, but be sure your board is tailored to fit the box and mounting blocks. An outline drawing of the cutout and mounting details of the board is shown in **fig. 4**. This should help you in getting a board cut to fit the box. Be careful not to run any wiring or conductors within about 5/16 inch of the board edges, since that's where the mounting blocks attach.

This versatile packaging technique adapts easily to other boxes. When carefully done, it results in a professional-looking enclosure providing the strong electrical bond between the board ground plane and the box in critical RF circuits. I've used this assembly

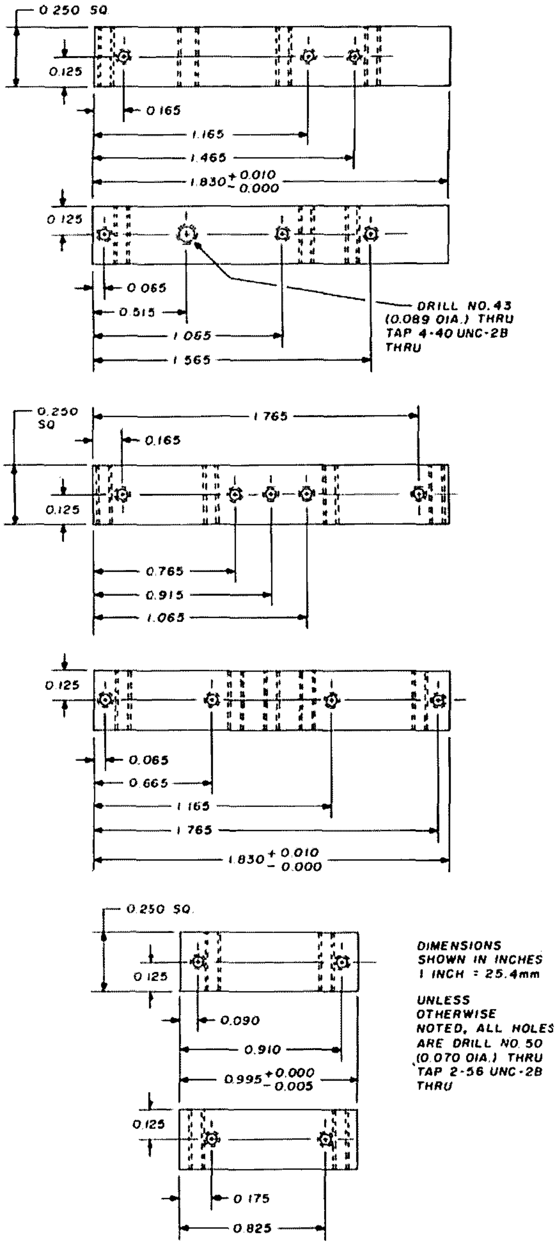

fig. 5. Side mounting block drilling details.

with excellent results for applications in excess of 1500 MHz; good luck on your future projects.

ham radio

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(See "Publisher's Log," April, 1984, page 6, for details.)

603-878-1441

✓ 143

cooling semiconductors
part 2:
blowers and fans

Heatsinks aren't the only way to dissipate excess heat

This article has been adapted, with permission, from *IC Voltage Regulator Sourcebook with Experiments*, by Vaughn D. Martin, published by TAB Books, Inc., and available from Ham Radio's Bookstore, Greenville, New Hampshire 03048 ($14 postpaid).

In Part One of this series we discussed the design and use of heatsinks in cooling electronic equipment. In this part, we'll examine the several other means by which excess heat can be prevented, monitored, and dissipated.

While heat sinks are without doubt the most familiar means of moving heat away from electronic equipment, they can be of limited use if the ambient temperature itself is high, or if the area surrounding the unit remains warm because heat is not evacuated quickly enough. If, for example, a power supply were enclosed in a poorly ventilated cabinet, a heat sink would be of little or no use; in such a situation, a fan or blower would be more effective.

While similar to heat sinks in principle, fans and blowers are significantly different in application and design. Fans (see **fig. 1**) generally employ propeller-like blades to move large volumes of low-pressure air. Blowers, on the other hand, usually consist of a revolving wheel that displaces air; they are therefore more efficient while operating near their maximum (non-moving) pressure (see **fig. 2**).

In cooling a poorly ventilated cabinet, pressurization of the enclosure by pumping filtered air *in* is vastly preferable to drawing air *out* because air pulled into the cabinet may contain particulate matter whose presence, over time, can compound the problem of temperature control by collecting in cabinet openings, such as those between panels or around doors, and blocking the exit of air.

Of the four methods of cooling electronic equipment housed in cabinets — forced-convection filtered air, air-to-air heat exchangers, air-to-water heat exchangers, and specially packaged air conditioning units — only the first (see **fig. 3**) is appropriate for Amateur Radio applications.

forced-air cooling

In designing the forced-convection filtered air-cooling of the interior of an equipment cabinet, five design guidelines should be followed: first, there should be no constrictions; the cross-sectional area of the air current should be at least as large as the intake. Second, the exhaust area must be located *downstream* from the heat-producing elements. Third, baffles work best when used to channel a small volume of air across an exceedingly hot component at a high velocity. Fourth, ducts may be used to maintain a more even cooling effect throughout the cabinet. If maintenance of an even temperature from the top to bottom of the cabinet is important, ducts should be located along the sides of the cabinet.

Finally, because blowers and especially fans can cause vibration, it's important that neoprene vibration isolators or similar devices be built into the fan if this is a concern. If you decide to use a fan, using the following two formulas will enable you to determine the required fan size:

$$\begin{aligned}\textit{volume of air at inlet} &= \frac{3.17 \times power \times 1.25}{T\ (°F)} \\ &= \frac{1.76 \times power \times 1.25}{T\ (°C)}\end{aligned}$$

By Vaughn D. Martin, 114 Lost Meadows, Cibolo, Texas 78108

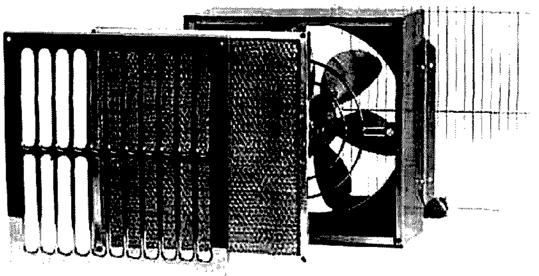


fig. 1. A typical fan used for ventilation purposes. (Photo courtesy McLean Engineering, Inc.)

The power to be dissipated is expressed in watts; the temperature is the average temperature rise within the cabinet, expressed in the first equation as degrees Fahrenheit and in the second equation as degrees Celsius. These two formulas as well as the nomograph shown in **fig. 4** have a built-in safety factor of 25 percent (note the 1.25 for 125 percent). If only a quick approximation is required, the nomograph may be used.

Let's work through a typical design example, assuming we are designing for a typical cabinet and using both the formula and the nomograph. Let's strike a line through the nomograph and see how close it comes to the exact calculated value.

Assume we have exactly 200 watts of power to be dissipated and can accept only a 10°F rise in temperature within the cabinet. The first equation for degrees Fahrenheit yields *[(3.17)(200 watts)(1.25)]/10 = 79.25 cubic feet per minute*. Note in **fig. 4** that the nomograph indicates approximately 80 cubic feet per minute — very close to 79.25, the value derived by actual calculation.

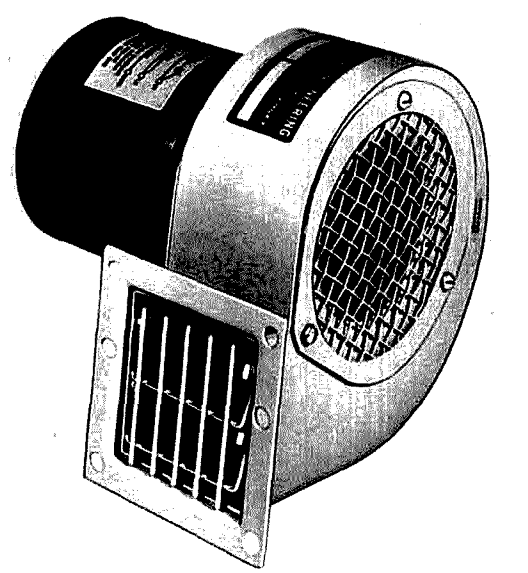


fig. 2. A typical blower along with a housing for blowers. (Photo courtesy McLean Engineering, Inc.)

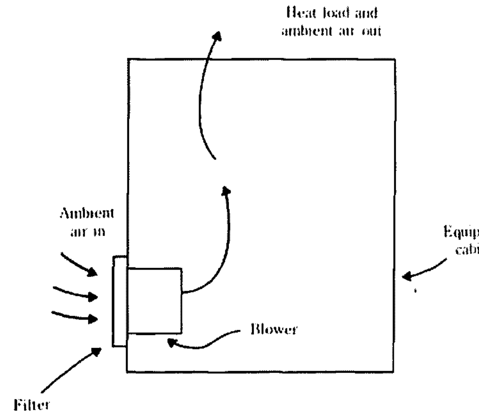


fig. 3. An illustration depicting forced-convention filtered air cooling. This technique uses ambient air to cool electronic control equipment. Incoming air is filtered before being blown throughout the equipment cabinet.

thermoelectric devices

In addition to forced air venting, thermoelectric devices such as Peltier and Thomson-Joule devices may be used to pull heat away from a specific small area in the cabinet or a critical semiconductor that must remain cool.

Four basic physical phenomena are associated with thermoelectric devices:

- The *Seebeck effect* is the EMF that exists when two dissimilar conductors are connected and have their junctions maintained at different temperatures. This is the basis of thermocouples.
- The *Thomson effect* is the heating or cooling effect that takes place in a homogeneous conductor when an electric current passes in the direction of the temperature gradient.
- The *Joule effect* occurs when an electric current passes through a conductor (which is isothermal or maintains the same temperature throughout) and generates heat — this is called "Joule heat."
- The *Peltier effect*, named in honor of Jean C.A. Peltier, who in 1834 discovered that the passing of electric current through the junction of two dissimilar materials causes a cooling effect when passed in one direction and a heating effect when the direction of current flow is reversed.

The Seebeck, Peltier, and Thomson effects are all

reversible phenomena. The Peltier effect is the one most useful in power supply design because of its ability to literally draw heat away from a semiconductor, although the Thomson effect also has some limited applications in these areas and will be discussed shortly.

Peltier cooling devices (**fig. 5**) may be attached directly to surfaces of heat-producing semiconductors. The effect of heat being drawn away in proportion to the current passing through the cooling device is what makes this device so effective, if somewhat expensive; a device large enough to handle a TO-3 power transistor is priced at about $15.00.

Another interesting thermoelectric device is the Joule-Thomson cooler, as shown in **fig. 6**. Described by its manufacturer as a "micro-miniature refrigerator," it cools a wide range of IR (infrared) and millimeter wave detectors down to 77°K, or Kelvin, which is the temperature of liquid nitrogen (N_2). Zero degrees Kelvin is *absolute zero* (where all molecular motion stops); this is equal to −273.15°C or −459.67°F. (One degree Kelvin equals in magnitude one degree Celsius, so to convert Kelvin to Celsius, just subtract 273.15 degrees from it. So *77°K = 77 − 273.15 = −196.15°C or −321.07°F.*)

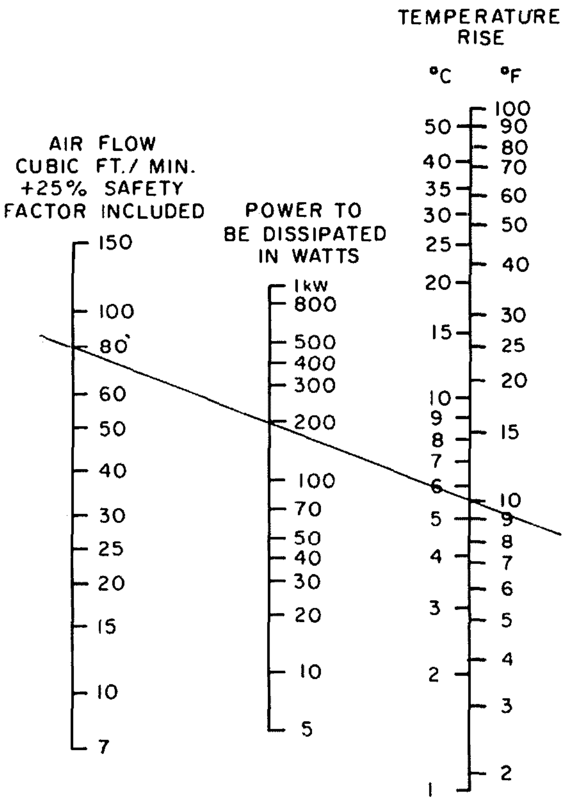


fig. 4. A nomogram for determining air flow requirements.

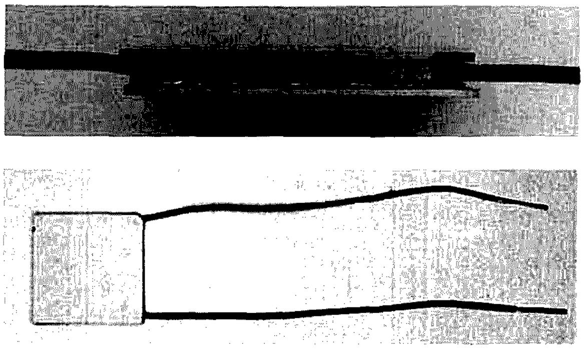


fig. 5. A Peltier cooling device (2 views).

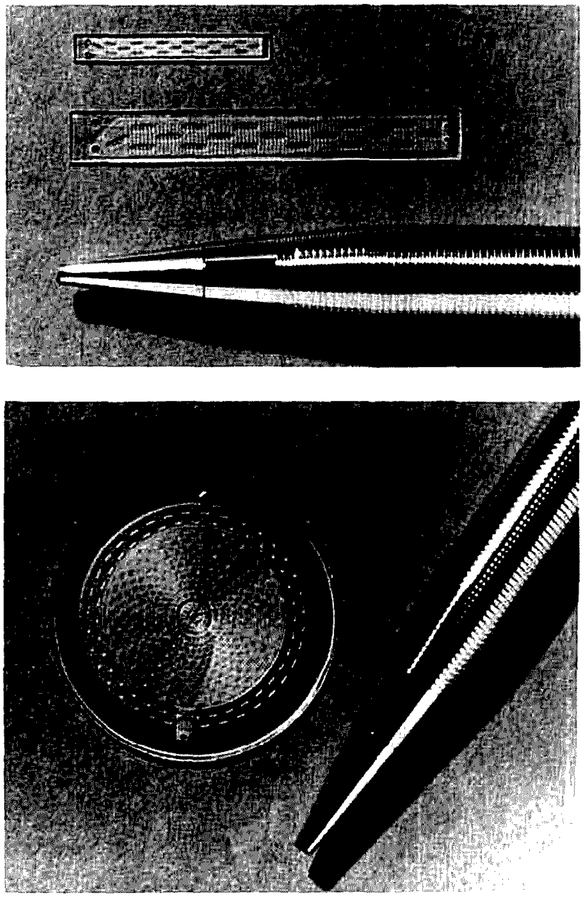


fig. 6. A Joule-Thompson cooler. (Photo courtesy MMR, Inc.)

These 1.5 cm (0.6 inch) in diameter devices can cool some detectors to 80°K (−193.15°C) in just one second.

piezoelectric fan

Another interesting cooling device is the piezoelectrically driven fan designed for PC board use. Manufactured by Piezo Electric Products,* this device (**fig. 7**)

*Piezo Electric Products, 212 Durham Avenue, Metuchen, New Jersey 08840.

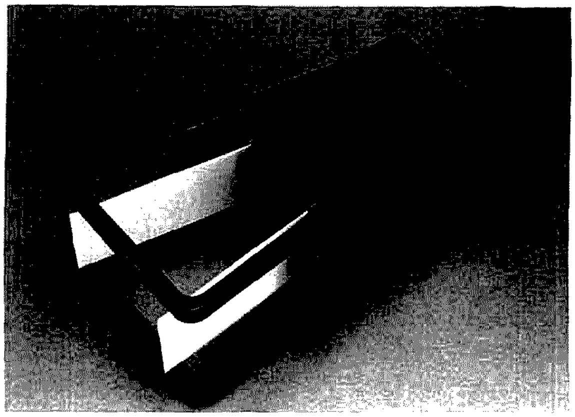
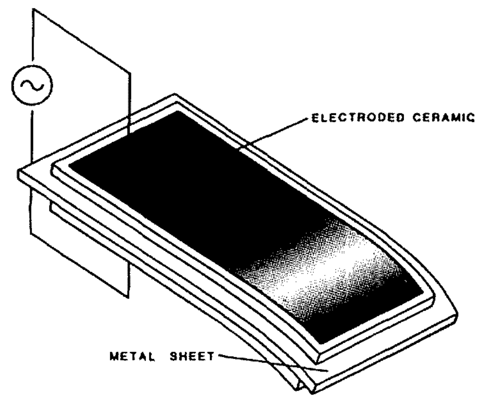


fig. 7. A piezoelectric fan. (Photo courtesy Piezo Electric Products, Inc.)

is available in either 50 or 60 Hz at 120 or 240 VAC. Used for spot cooling, it consumes about 1/15 of the power of a fan, *yet accomplishes the same cooling effect*. Its reliability is very high because of flexible metal strap blades laminated to thin-sheet piezoceramic elements. The mechanical vibrations that result when an AC voltage is applied across the piezoelectric element in the bender assembly cause the attached plastic blades to flap. This principle is illustrated in **fig. 8**. The highly "focused" air streams are responsible for the unit's exceptional efficiency; the fan moves 20 cubic feet/minute of air while using just 0.36 watts.



WHEN VOLTAGE IS APPLIED, A BILAMINATE BENDER DEFLECTS AS ONE CERAMIC SHEET EXPANDS WHILE THE OTHER CONTRACTS.

fig. 8. An illustration of the principle behind the piezoelectric fan.

the heat pipe

In size, heat pipes range from very small — about the size of a pipe cleaner (**fig. 9**) — to very large, as in furnace linings and factory smokestacks. The heat pipe's claim to fame is its unique ability to equally and uniformly distribute heat and thus equalize the differences between the hottest and coolest objects (usually ICs and ambient air, respectively). The smallest heat pipes fit nicely beneath and between the two adjacent rows of IC pins.

Heat pipes are metallic, sealed, self-contained units (**fig. 10**) with a thin outer wall or shell of copper which surrounds an internal wick saturated with a working fluid. As one end of the heat pipe (the evaporator end) is heated, the fluid within vaporizes and travels to the other end (the condenser end). As the vapor condenses, heat is given off and the vaporized liquid condenses and travels back, via capillary action, to the evaporator end. The end result is an efficient self-contained heat exchange system with the following advantages:

- The ability to transfer over 1000 times more heat than a copper rod of the same size and weight.
- Virtually absolutely uniform heat transfer and distribution throughout the heat pipe with less than 0.1 degree C drop or heat gradient from one end of the pipe to the other.

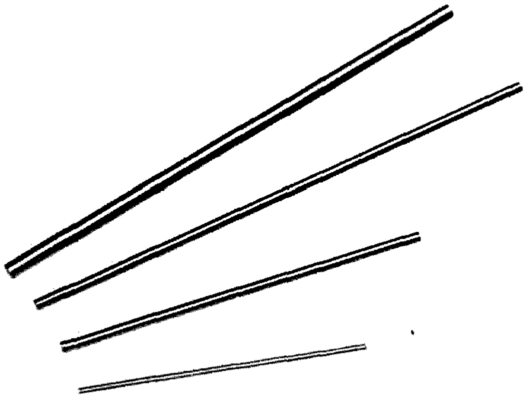


fig. 9. Typical small heat pipes. (Photo courtesy Noren Products, Inc.)

- Lightweight and compact.
- No moving parts, no maintenance or external power required.
- No noise — either electrical or audible.
- Capability of being fused to an existing heat sink (**fig. 11**).

If, for example, you had an LED display composed of a row of adjacent DIPs, and very poor or no ventilation, as in **fig. 12**, you wouldn't be able to use a slip-on DIP heat sink such as the one shown in **fig. 13**; slipping this heat sink over the DIP displays would obscure the front of the display, rendering it useless. In

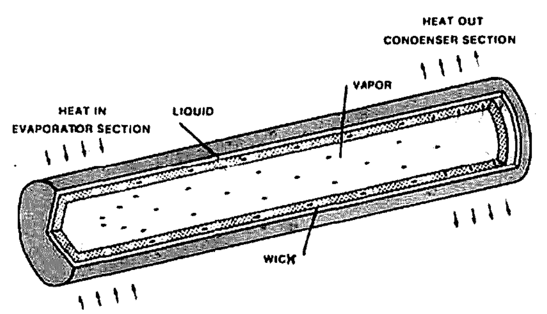


fig. 10. An illustration of the heat pipe's operating principle.

applications such as this, a heat pipe is ideal. When it's necessary to vent heat away *after* it has been transferred and evenly distributed, a radiator, as shown in fig. 14, may be attached to the heat pipe.

monitoring temperature

After heat sinks, heat pipes, and either a fan and/or a Peltier or Thomson-Joule device are added, the temperature should be within acceptable limits, and temperature monitoring should begin. A number of companies manufacture a crayon-like pen that can be used to apply a heat-sensitive substance to the surface of a semiconductor whose temperature is to be monitored (fig. 15). There are also TO-3, TO-66, and DIP stick-on temperature indicators (fig. 16) and other heat-sensitive products such as paints and tablets (fig. 17) that change colors in direct relation to the temperature they monitor; as the temperature changes, so does the color of the temperature indicator. These indicators are available in two types. The more traditional throwaway types change color as temperature rises, but if the temperature falls again, cannot change back to a color corresponding to this lower temperature. This unfortunate "ratchet" effect can be avoided by use of an LCD temperature indicator. Instead of the more traditional nematic crystals, these indicators use cholesteric crystals — produced in the cholesterol in lambs wool and cuttlefish[1] — and cost about four dollars each, feature as many as seven colors, and have 5-degree C temperature increments as standard. There are larger indicators based on the same principle that can cover whole power supplies or entire printed circuit boards.

protection devices

In order for overheating to be controlled, it must first be sensed. The visual methods just described (shown in fig. 15 through 17) are one form of indicator; however, in unattended applications, electrical parameters, as well as heat, must be sensed and controlled. In a typical power supply, there is a fuse at the transformer, but other protection can be provided for individual component parts within the system.

A PTC (positive temperature coefficient) thermistor can help. These devices are very inexpensive — about \$1 — and are two-lead devices that sense ambient heat by having the body or encapsulating cover experience heating, causing its resistance to rise by an order of magnitude (a factor of ten) for each 10 degree C rise in temperature above its trip point. This trip point is different per unit and can be specified, but a typical trip point is about 100 degrees C. Once this elevated

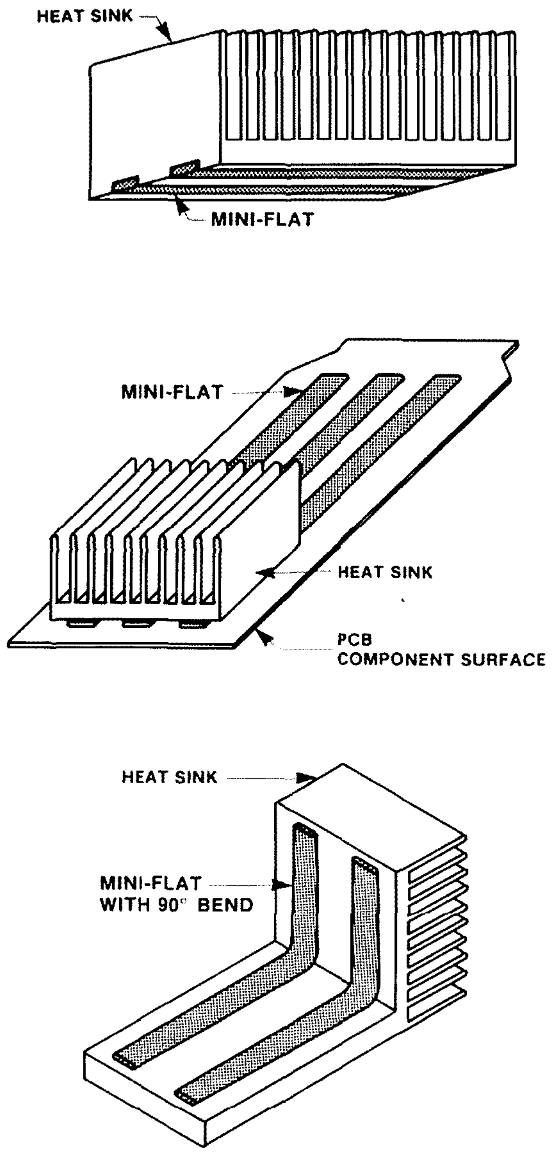


fig. 11. Various schemes of fusing heat pipes into existing heat sinks to increase their effectiveness.

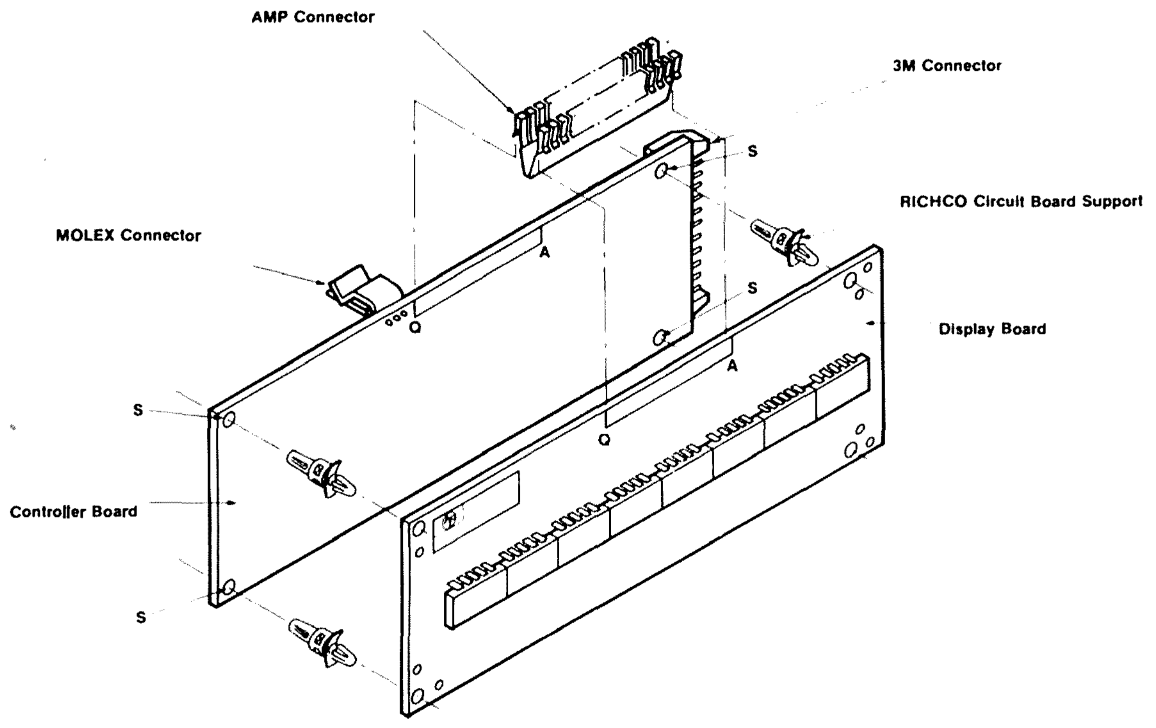


fig. 12. An ideal candidate for a heat pipe application, a poorly vented LED display.

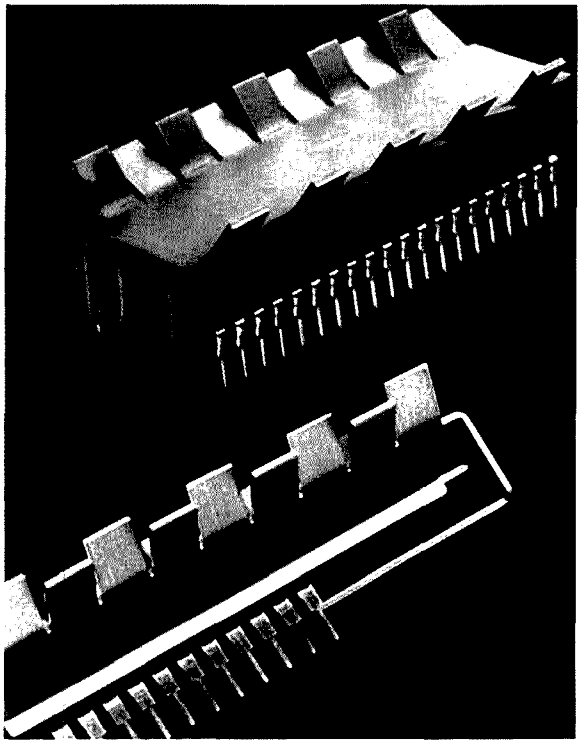


fig. 13. A DIP slip-on heat sink, unfortunately not useful in an application such as with the display in *fig. 12*. (Photo courtesy Aavid Engineering, Inc.)

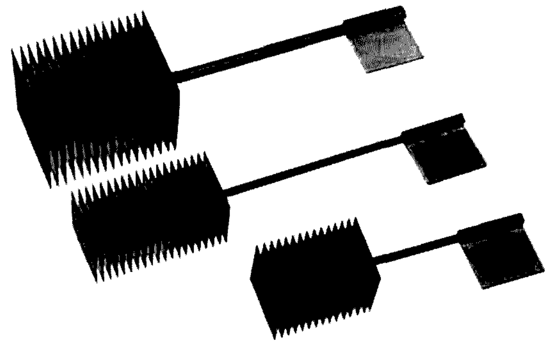


fig. 14. A "radiator" attached to a heat pipe for more effective heat radiation. (Photo courtesy Noren Products, Inc.)

fig. 15. Heat sensitive crayons used to visually detect temperature changes. (Photo courtesy Tempil Division, Big Three Industries, Inc.)

temperature is reached, though, automatic reset by merely removing power to the PTC thermistor is not possible. An actual cooling of the PTC thermistor to below its trip temperature must first occur before re-

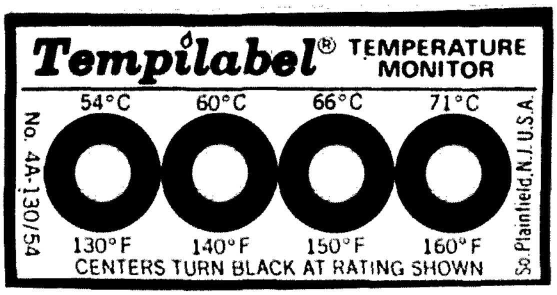


fig. 16. Heat sensitive sticky backed peel-off temperature indicators.

fig. 17. A whole product line of visual temperature indicators. (Photo courtesy Omega Engineering, Inc.)

setting is possible. The internal heating effect of the thermistor itself is inadequate to cause it to reach the trip temperature. This allows internal heating to have only a very small effect on the resistance of this heat sensitive resistor. It also allows the surface that is producing heat, usually flush with the PTC thermistor, to control the device's resistance. **Fig. 18** is a graph showing how this internal heating effect is minimized by a PTC.

summary

Though useful for the applications intended, heat sinks are not the only means of dissipating heat, especially within an equipment cabinet. Blowers and

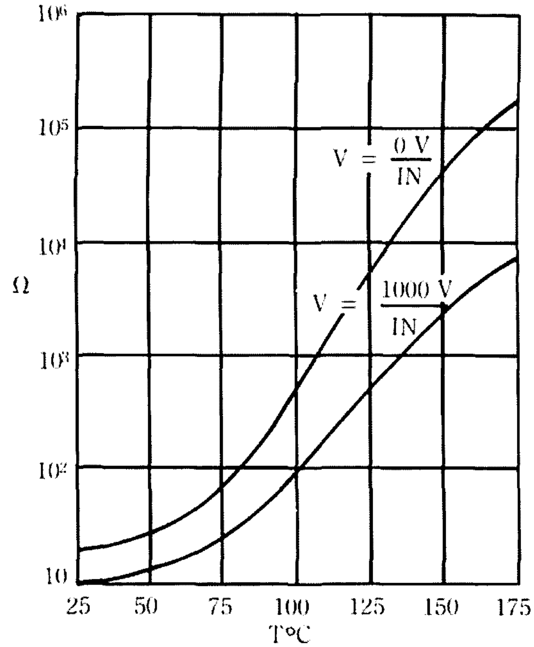


fig. 18. A graph showing how a PTC (positive temperature coefficient thermistor) can overcome the effects of internal device heat generation. This graph demonstrates the sensitivity of resistance of PTC material to applied voltage.

fans can be selected and sized to maintain specified temperature rises for a given power dissipation requirement. On a smaller scale, thermoelectric devices exhibiting the Seebeck, Thomson, Joule, or Peltier effect can be used to cool individual semiconductors. Heat pipes are also very effective in uniformly transferring heat from one location to another. The effectiveness of these devices can be determined by using temperature indicators in the form of stick-ons or paints. Overheating protection devices that generate voltages in a feedback loop can be used to reduce the original cause of heat buildup.

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ham radio

DX FORECASTER

Garth Stonehocker, KØRYW

summer noise

Summertime DX really isn't too bad on the higher frequency bands, 6-30 meters. Between sporadic-E short-skip openings near noontime and the bands staying open longer, enough DX fun is still available to keep us happy all day and well into the evening. The bands hardest hit by summertime problems are 80 and 160 meters of the lower-frequency bands. Their problem is summer thunderstorm noise — QRN.

Thunderstorm noise propagated from the equatorial regions increases the overall average noise level of the lower-frequency HF bands. At any given moment, 3600 storms are in progress around the world. That's a lot of QRN! Some, of course, are inevitably nearby.

Air-mass thunderstorms, which form only over land, build up from the hot summer sun's heating the ground and the air above it. They form in the afternoon if the humidity is above 50 percent, and last into the night before cooling off enough to dissipate. Unlike spring and fall frontal passage thunderstorms, which simply pass by your QTH, air-mass thunderstorms stay around for days until they release their moisture in the form of rain. During the evening DXing hours, the air-mass thunderstorm QRN more or less limits the usefulness of these band's signals to local ragchewing, and rule out weak-signal DX.

So how do you get some DXing in on these bands? Most operators switch operating hours, giving up evenings in favor of the pre-dawn hours of early morning. By this time, the thunderstorms have dissipated locally and are dissipating on paths to the west. This is a cool, comfortable time of the day to be up and around. Good luck, early bird!

last-minute forecast

August is the last full month of summertime DX conditions, characterized mainly by sporadic-E short skip and longer daylight for high-frequency band DX operating. These higher frequencies are forecast to be best during the first two weeks of the month because of an expected high solar flux. The lower frequency bands, 30-160 meters, should improve during the last two weeks of the month because of lower noise and lower absorption of the signal's energy. This trend will become even more apparent next month.

The moon's perigee will occur on the 27th, with a full moon on the 11th. The Perseids meteor shower occurs from the 10th to 14th, with its maximum on the 11th and 12th, with better than fifty meteors per hour. This is an excellent shower.

band-by-band summary

Six-meter paths will open for a half hour to a couple of hours on some days around local noon. Sporadic-E propagation will make this short-skip path possible out to nearly 1200 miles (2000 km) per hop.

Ten, fifteen, twenty, and thirty meters will support DX propagation from most areas of the world during daylight and into the evening with long-skip out to 2000 miles (3500 km) per hop. Sporadic-E short skip will also be available on many days for several hours near local noon. The direction of propagation will follow the sun across the sky: morning to the east, south at midday, and toward the west in the evening. Daylight is still long, providing many hours of good DXing.

Thirty, forty, eighty, and one-sixty meters are the nighttime DXer's bands. On many nights 30 and 40 meters will be the only usable bands because of thunderstorm QRN. Try the pre-dawn hours for less QRN. The direction of propagation follows the darkness path across the sky: evening to the east, south around midnight, and toward the west in the pre-dawn hours. Distances will decrease to 1000 miles (1600 km) for skip on these bands. Sporadic-E openings will be most frequently observed around sunrise and sunset. These may be the only signals getting through the noise in the evening.

ham radio

WESTERN USA

GMT	PDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	5:00	20	20	15	10	15	10	10	20
0100	6:00	20	20	20*	10	15	10	10	20
0200	7:00	20	20	20	10	20	10	10	20
0300	8:00	20	20	20	10	20	10	10	20*
0400	9:00	20	20	20	10	20	15	10	20*
0500	10:00	20	20	20	10	20	15	10	20
0600	11:00	20	*20*	20	15	20	15	15	20
0700	12:00	20	*20*	*20*	15	*30*	15	15	20
0800	1:00	20	*30*	*20*	*15*	*30*	20	15	20
0900	2:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	20
1000	3:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	20
1100	4:00	*20*	30	*20*	*20*	*30*	*20*	*20*	*30*
1200	5:00	*20*	30	*20*	20	*30*	*20*	*20*	*30*
1300	6:00	*20*	30	15	20	*30*	*20*	*20*	*30*
1400	7:00	30*	20	15	20	*30*	*20*	*20*	*30*
1500	8:00	30*	20	15	20	20	*20*	*20*	*30*
1600	9:00	30*	20	15	20	20	15	20	*30*
1700	10:00	30*	20	15	20	20	15	20	20
1800	11:00	30*	20	10	20	20	15	20	20
1900	12:00	30*	20	10	15	20	15	15	20
2000	1:00	20	20	10	15	20	15	15	20
2100	2:00	20	20	10	15	20	10	15	20
2200	3:00	20	20	15	15	20*	10	15	20
2300	4:00	20	20	15	15	15	10	15	20

AUGUST

MID USA

MDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN	CDT
6:00	20	20	15	15	15	10	10	20	7:00
7:00	20	20	20*	15	20	10	10	20	8:00
8:00	20	20	20	20	20	10	10	20	9:00
9:00	20	20	20	20	20	10	10	20	10:00
10:00	20	20	20	20	20	15	10	20	11:00
11:00	20	20	20	20	20	15	15	20	12:00
12:00	20	*20*	20	20	20	15	15	20	1:00
1:00	*20*	*30*	*20*	20	*30*	20	15	20	2:00
2:00	*20*	*30*	*20*	*20*	*30*	20	15	20	3:00
3:00	*20*	*30*	*20*	*20*	*30*	20	*20*	*20*	4:00
4:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	*30*	5:00
5:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	*30*	6:00
6:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	*30*	7:00
7:00	*20*	20	15	15	*30*	*20*	*20*	*30*	8:00
8:00	20	20	15	15	*30*	*20*	*20*	*30*	9:00
9:00	20	20	15	15	20	*20*	*20*	*30*	10:00
10:00	20	20	15	15	20	15	20	*30*	11:00
11:00	20	20	15	10	20	15	20	20	12:00
12:00	20	20	10	10	20	15	20	20	1:00
1:00	30*	20	10	10	20	15	20	20	2:00
2:00	20	20	10	10	20	15	15	20	3:00
3:00	20	20	10	10	20	10	15	20	4:00
4:00	20	20	15	15	20	10	15	20	5:00
5:00	20	20	15	15	15	10	15	20	6:00

EASTERN USA

EDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE CARIBBEAN S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
8:00	20	20	15	15	20	10	10	20
9:00	20	20	20*	20	20	10	10	20
10:00	20	20	20	20	20	10	10	20
11:00	20	20	20	20	20	15*	10	20
12:00	20	20	20	20	20	15	10	20
1:00	20	30*	20	20	20	15	15	20
2:00	*20*	*30*	20	20	30	15	15	*20*
3:00	*20*	*30*	20	20	30	20	15	*20*
4:00	*20*	*30*	*20*	*20*	30	20	20	*20*
5:00	*20*	*30*	*20*	*20*	*30*	20	*20*	*20*
6:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	*30*
7:00	*20*	*30*	*20*	*15*	*30*	*20*	*20*	*30*
8:00	*20*	*30**	15	*15*	*30*	*20*	*20*	*30*
9:00	*20*	20	15	15	*30*	*20*	*20*	*30*
10:00	20	20	15	15	*30*	*20*	*20*	*30*
11:00	20	20	15	15	*30*	*20*	*20*	*30*
12:00	20	20	15	10	20	15	20	*30*
1:00	20	20	10	10	20	15	20	20
2:00	20	20	10	10	20	15	20	20
3:00	20	20	10	10	20	15	15	20
4:00	20	20	10	10	20	15*	15	20
5:00	20	20	10	10	20	10	15	20
6:00	20	20	15	15	20	10	15	20
7:00	20	20	15	15	20	10	15	20

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during 'normal' hours.
*Look at next higher band for possible openings.

fundamentals of grayline propagation

Signals don't always follow "normal" paths

The decline of the sunspot cycle has been accompanied by a corresponding decrease in DX opportunities on the higher frequency bands. Ten meters is becoming very sporadic in nature, as F-layer path openings continue to decrease in both number and duration. Within a few years, at the bottom of the solar cycle, the only activity remaining on 10 meters will be via sporadic-E propagation. The 15-meter band is not much better; as 10 meters goes downhill, so goes 15: this band will also continue to decline, offering fewer and poorer openings of much shorter duration.

As the levels of solar flux continue to drop, there will be an increasing number of days when the only (high HF band) activity will be found on 20 meters. Even this band won't offer the same quality openings of several years past. During many of the years to come, it will be a strictly daylight DX band with only scattered openings to various parts of the world.

With the knowledge that much of the DX has migrated to 40 and 80 meters, and believing that high power is required on these bands, many Amateurs feel that there's no place left for the "barefoot" DXer. After all, even though the low power operator can compensate for the lack of a kilowatt with a good beam and tower combination on the higher frequency bands, how many of us have the money and/or acreage for beams or phased arrays on 80 or 40 meters? Today, many low power operators will attempt to work DX on these bands only during contests, believing that their signals will never be heard at any other time.

try grayline

Is there really no place left for the "barefoot" DXer? Of course not: nature has provided us with a mode of propagation on these low bands that will allow DX operation without the use of maximum power or costly low-band directive arrays. This type of propagation — called "grayline" or "shadow edge" propagation — takes place along that line dividing the daylight and darkness hemispheres of the planet, functions on both 80 and 40 meters, and offers outstanding conditions during all phases of the sunspot cycle. What's more, it's at its *best* during the years of low solar activity!

The line that divides the region of the Earth in daylight from the nighttime side is called the *terminator*. The terminator is not a distinct division, but instead a gray, "twilight" band approximately 1000 miles in width and stretching completely around the Earth. The time at any point along this band will be that of local sunrise or local sunset.

As shown in **fig. 1**, the position of the terminator changes as the position of the Earth changes with respect to the Sun. (This familiar phenomenon causes the change in seasons of the year and is due entirely to the tilt of the Earth's rotational axis. Since this axis is not perpendicular to the plane of the ecliptic, but rather tilted 23 degrees, 27 minutes, the terminator swings back and forth through an arc of just under 47 degrees during the course of a year. The northerly and southerly limits of this motion are delineated by the Antarctic and Arctic circles, respectively.)

The absorptive and refractive characteristics of the ionosphere are directly related to the angle of solar radiation. Normally, the higher this angle (the more nearly the Sun is to vertical), the more enhanced these characteristics become. Conversely, at very low angles of radiation, the absorption characteristics change much more quickly than do the refractive characteris-

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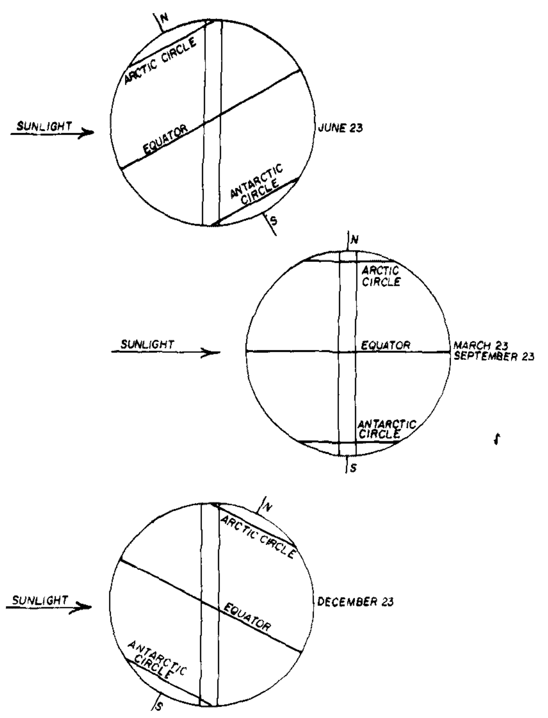

fig. 1. Motion of the terminator with the changing seasons.

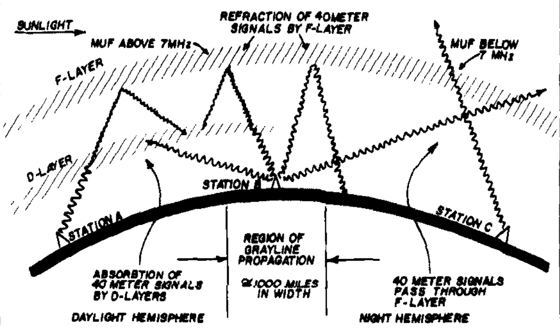

fig. 2. The mechanics of grayline propagation viewed at right angles to the Earth's terminator.

tics. Thus, the ionosphere in the twilight regions of Earth is in a state of flux and rapidly changing conditions prevail.

Because of the rapidly shifting properties of the ionosphere within this region, propagation of certain frequencies along the terminator is extremely efficient. **Fig. 2** represents a cross-sectional view of the terminator and the ionosphere in and around this region. Within the daylight hemisphere, both the D-layer and F-layer are near their peak levels of ionization. High levels of ionization cause the D-layer to become an effective absorber of radio frequency energy within the frequency range of 300 kHz to approximately 10 MHz. While the F-layer is capable of refracting radio signals within this range of frequencies, the high levels of absorption within the D-layer prevent these signals from reaching the F-layer. This explains why the 160, 80, and 40-meter bands are essentially nighttime bands except for local or regional communications.

visualizing grayline

To visualize how grayline propagation works, examine station "A" in **fig. 2**. Located in the daylight hemisphere, its 40-meter signal is severely attenuated as it passes through the D-layer. That signal which remains is refracted by the F-layer, only to be further attenuated as it passes through the D-layer on its return path. Some RF energy will, in fact, reach the surface, but it will be well below the threshold of detectability in Amateur receivers. The only way to avoid this dilemma would be to increase power to levels well above those allowed in the Amateur service or to increase integration time of receiver detection systems well beyond that which is required for either CW or SSB.

Station "C," in **fig. 2**, suffers from the opposite problem. The D-layer is nonexistent due to the lack of solar radiation, but because of changes in the F-layer, the MUF (Maximum Usable Frequency) has dropped below 7 MHz. As a consequence, the 40-meter signal radiated by station "C" penetrates the ionosphere and is lost.

Now look at station "B," located in the terminator region on the Earth's surface. Signals radiated toward the daylight region are as severely attenuated as those of station "A," while the signals radiated toward the nighttime region pass through the ionosphere, as are those of station "C." It would seem that station "B" is in no more a favored position than those other two stations. **Fig. 3** illustrates what would happen in actual operation. Signals radiated parallel to the terminator are propagated for long distances with little attenuation.

Because the terminator, shown in **fig. 3**, extends completely around the Earth, it is possible to work into any area of the world within this region. This situation is most noticeable on 80 and 40 meters and, to a lesser degree, on 20, 30, and 160 meters. Since these low-band signals suffer little attenuation on this path, long distance communications are possible with low power levels and ordinary non-directional antennas, such as dipoles and verticals.

There are, unfortunately, some areas of the world you'll never be able to work using grayline; this is because any two selected areas on the surface of the Earth do not necessarily share a common terminator. If the Earth's rotational axis were tilted 45 degrees from the plane of the ecliptic, then, at some time or other

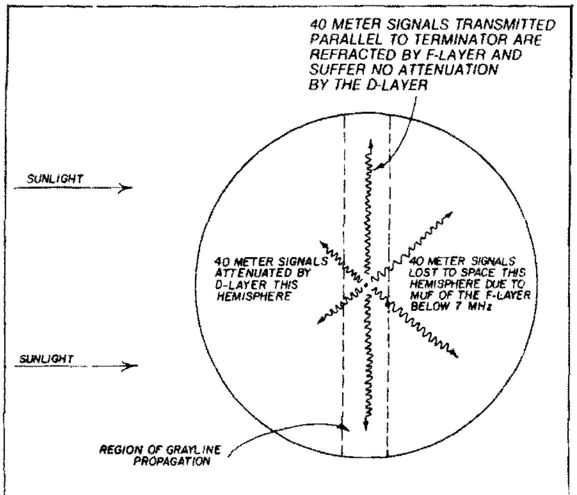

fig. 3. Radio wave propagation along the terminator.

during the year, your QTH would be on the terminator with every other spot on the Earth's surface. In addition, the position of the terminator and hence the path of grayline propagation is not stationary, but instead moves with the changing seasons because of the yearly movement of the Earth's rotational axis with respect to the Sun.

Two other factors enter into grayline DXing. Remember that local time will vary all along the length of the terminator. Since most Amateurs are more active in the local afternoon and early evening, rather than at dawn, time your operating to coincide with that time when *they* are most active. Thus, to work hams on the other side of the world with this mode, the best time would be between 30 minutes before to 30 minutes after your local sunrise. In addition, remember that the position of the terminator at sunrise is not the same as its position at sunset, local time.

Figs. 4, 5, and **6** show the positions of the terminator at various times throughout the year. Notice that the only time the sunrise and sunset paths coincide is twice a year, during the spring and fall equinoxes. *The information shown in these three charts is valid only for a station located in the Pacific Northwest and will be different for stations located elsewhere.** It is apparent that a station in Seattle could work half of Africa, Europe, much of Asia, parts of South America, and much of the South Pacific during the course of a year using grayline propagation on 80 and 40 meters.

determining grayline paths

There are several methods used to determine grayline paths for your location. The easiest, but most cumbersome, involves the use of a globe and a piece of thin cardboard. *Any* globe will do, even the type sold as piggybanks in the local five-and-dime. Cut a circular hole in the center of the cardboard just big enough to allow it to slip over the globe. It can then be positioned on the globe to represent the grayline path during any particular day of the year. This method can also be used to determine the grayline path for

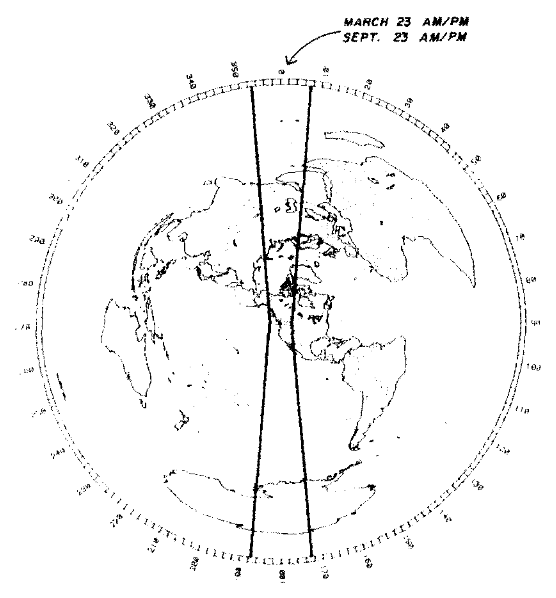

fig. 4. Grayline paths during the equinoxes. (Azimuthal-equidistant maps prepared by Bill Johnston, N5KR.)

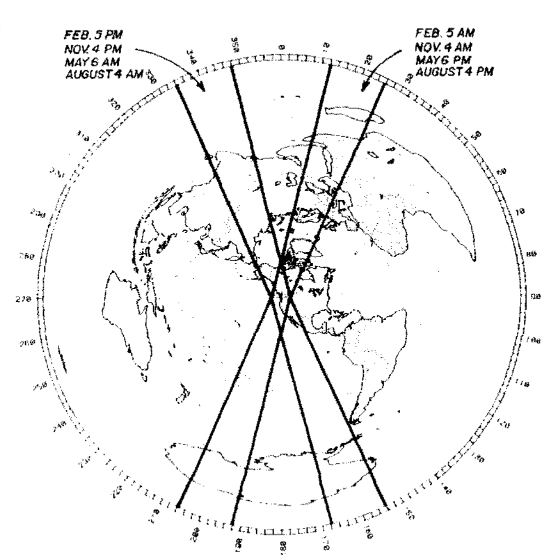

fig. 5. Grayline paths during several non-equinox periods. (Azimuthal-equidistant maps prepared by Bill Johnston, N5KR.)

*Custom-made azimuthal-equidistant (great circle) maps centered on your own location are available from Bill Johnston, N5KR, 1808 Pomona Drive, Las Cruces, New Mexico 88001.

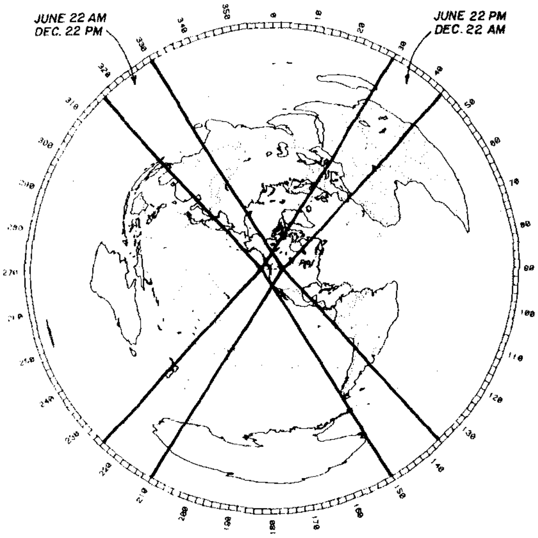

fig. 6. Grayline paths during the longest and shortest days of the year. (Azimuthal-equidistant maps prepared by ***Bill*** **Johnston, N5KR.)**

selected times during the year and plot the results on a map as was done in **figs. 4, 5,** and **6.** (These base maps are available through a number of sources, most of which advertise in the various Amateur magazines.) By taking the time to devise three sets of maps, the approximate grayline path for any day of the year can be quickly and easily determined.

Perhaps the most convenient method of figuring grayline paths is through the use of "The DX Edge," slide rule-type device with a map of the world printed on one side and equipped with a series of overlays representing the grayline paths for each month of the year.* By sliding the appropriate overlay back and forth along the map, the path of the terminator across the Earth's surface is easily determined.

Regardless of the method used, figuring and using grayline propagation paths will enhance your ability to work low-band DX. Of course, the old adage "The higher the antenna, the better" is still true. But the fact that you don't have a directional antenna array or a 2000 watt amplifier is no excuse for not snagging the rare ones. Patience, perserverance, and a knowledge of grayline propagation will catch all the 80 and 40-meter DX you can handle, and you won't have to stay up until two o'clock in the morning to do it.

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Jacobs, George, W3ASK, and Cohen, Theodore, N4XX, *The Shortwave Propagation Handbook*, CQ Publishing, Inc., Hicksville, New York, 1982.

ARRL Ham Radio Operating Guide, American Radio Relay League, Newington, Connecticut, 1976.

ham radio

**DX Edge* is available from Ham Radio's Bookstore, Greenville, New Hampshire 03048 *($16.95 plus $1.50 postage and handling).*

DXing by computer

Program determines grayline, distance/azimuth, and MUF/FOT

You're an avid contester who needs an elusive KX6 multiplier on 10 and 15 meters during the VK/ZL contest in October; when should you spend time looking for this rare prefix? Or perhaps you don't care too much about DXing, but are interested in talking with a friend in Pakistan; which band and at what time should you expect propagation to be at its best? Or maybe you've heard about grayline propagation, but don't know how to tell what countries are on the grayline or when to try to make contact with them. What's the beam heading to South Africa? How far is the long path distance? If you do make contact with your friend in Australia, how good will the band conditions be? Should you use CW or SSB? It's questions such as these that suggest that the only way to manage the variety and volume of information available to Amateurs is by computer.

The user-friendly program described in this article consolidates the best features of many programs already on the market. It makes available, for the first time, a means of accurately determining the sunrise and sunset grayline for any location on the earth and for any desired day during a year.

A summary of the program's features is shown below, and a short discussion of the outputs follows. In addition, several examples of how to use the program are demonstrated by answering some of the questions raised at the beginning of the article.

User inputs (data entered) include the date, base location, the name of any of 433 target locations (countries, states, or provinces), DX location (target), 10.7 cm solar flux, geomagnetic A factor, and the grayline desired and its width.

Printed program outputs include the grayline output and screen output. Computer inputs include the following:

Base sunrise/sunset times. This is normally the user's operating location sunrise/sunset times given in UTC, or more popularly, GMT. The algorithm used to calculate the sunrise and sunset times has been extensively compared to world and nautical almanacs *and no error greater than 1 minute has been found.*

Target sunrise/sunset times. If the target selected is a single point — that is, one specific location — the output will give only the target average sunrise/sunset time. If you have chosen a country for your target location, and the country is large enough to have significantly different sunrise/sunset times, then the output will include a *minimum and maximum* sunrise and sunset time. This output can be used to determine the coincident darkness and daylight between you and the desired target. In general, look for 10 and 15-meter openings during daylight coincidences (except in years of very high sunspot activity). Look for openings on

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40, 80, and 160 meters during darkness coincidences. *Twenty meters is normally* open most days and nearly through the night to some areas of the world.

Long/short path output. The program computes the long path and short path great circle distance in nautical miles *(1 nautical mile = 1.15 statute miles = 1.85 kilometers)* and also computes the long and short path compass beam headings between the user's operating location and the desired target location.

Sunspot number/quality factor. The program determines the sunspot number based on the 10.7 cm solar flux. The quality factor comes from the August, 1982, issue of *ham radio* and is an empirical means of describing the quality of a QSO, with 0 being poor and 9 being very good.[1]

MUF/FOT. The MUF or Maximum Usable Frequency is a definite boundary that the HF user cannot overcome with power, antenna, or other mechanical means. The MUF is the range of frequencies that can be used before radio waves penetrate the F layer and not return to earth. *Because of minimum signal absorption at this frequency, the MUF gives the greatest signal strength at the receiving point for a given transmitted power.* Fifty percent of the month the highest propagated frequency achieves this value — MUF. Ninety percent of the month, the highest propagated frequency will achieve the value known as FOT (Frequency of Optimum Transmission). FOT is also included in the program and uses the same algorithm as was used to do MUF. This section of the program makes use of information provided by Robert Rose, K6GKU, on MINIMUF.[2]

Grayline. Observation has shown that enhanced conditions for working DX often occur at sunrise and sunset. This is particularly true when both the base location and the desired target location are undergoing sunrise or sunset at the same time. This section of the program provides an easy means of determining who else in the world is experiencing sunrise or sunset at the same time that you are. The grayline is the great circle path that describes the line drawn on the earth that separates the half that is in daylight from the half in darkness. The width of the grayline is a variable that is determined by the user; for most applications, a time window of ± ½ hour is sufficient to determine which countries can be found. This corresponds to a grayline width of 15 degrees.

The two questions that are answered by this part of the program are first, for any day, when will my sunrise/sunset occur? And second, what will be the bearing of the grayline, and what targets will lie along it?

table 1. Program output display showing information and format available.

```
[1] JAN 1                                  PATH
          LAT    LONG                 SHORT   LONG
[2] BASE  00:00  000:00       DIST      0     21600
[3] TGT   00:00  000:00       HEAD      0       180

          SUNRISE   SUNSET
BASE       05:59     18:07    [4]   GEO-MAG 0
TGT-AVE    05:59     18:07    [4]   10.7 cm 64
TGT-MIN                             SUNSPOT 0
TGT-MAX                             Q-FACT 4.1

[5] MAXIMUM USABLE FREQUENCY (FOT [6])
00        05        10        15        20
01        06        11        16        21
02        07        12        17        22
03        08        13        18        23
04        09        14        19        24

ENTER 1-6, (G)-LINE, (P)RINT
```

As you can see from the list above, the number of inputs needed is really very small and readily available from just a few sources. Using any world atlas, the latitude and longitude of the base location — this is *where you are* — can be determined. Again, using the world atlas, any specific target location — this is *where you want to go* — can be obtained. The data base of 433 unique radio locations stored in the program eliminates the need to know or look up the latitude or longitude of a specific country; all you need to know is the name of the country. The 10.7 cm solar flux and the geomagnetic A index can be obtained from WWV. Listen to any of the assigned frequencies you are able to receive, (2.5, 5, 10, 15 and 20 MHz) at 18 minutes past the hour or use the hotline telephone number (303-497-3235). The program uses the 10.7 cm flux to calculate the sunspot number, and the geomagnetic *A* index to determine the quality factor. The two final items that the program needs are the date for which you wish to calculate the program outputs, and the width of the grayline.

using the program

After the program is loaded, the monitor will display information similar to that shown in **table 1**. Three distinct sections are seen on the screen and in **table 1**: data, coordinates, distances and headings; sunrise, sunset, and the geophysical information; and MUF and FOT. The numbers in brackets indicate where the user must input data to the program. All input routines are user-friendly and have built-in error detection to prevent data that is out of range: for example, February 31 or a 10.7 cm solar flux of 0 (minimum solar

table 2. Kansas to Marshall Island path parameters.

OCT 15

	LAT	LONG		PATH SHORT	PATH LONG
BASE	37:00	100:00	DIST	4802	16798
			HEAD	275	95

TARGET NAME: MARSHALL ISLANDS
TARGET PREFIX: KX

	SUNRISE	SUNSET	
BASE	12:46	00:06	GEO-MAG 8.0
TGT-AVE	18:08	06:04	10.7 cm 133.0
TGT-MIN	18:08	06:04	SUNSPOT 86
TGT-MAX	18:08	06:04	Q-FACT 4.9

MAXIMUM USABLE FREQUENCY

0	30.9	5	16.8	10	13.1	15	14.4	20	29.0
1	28.6	6	15.7	11	12.7	16	13.7	21	31.1
2	25.0	7	14.9	12	12.4	17	17.6	22	32.6
3	19.5	8	14.1	13	12.2	18	22.7	23	32.4
4	18.0	9	13.6	14	15.1	19	26.3	24	30.9

FREQUENCY OF OPTIMUM TRANSMISSION

0	26.2	5	14.2	10	11.1	15	12.2	20	24.7
1	24.3	6	13.4	11	10.8	16	11.7	21	26.4
2	21.3	7	12.6	12	10.6	17	14.9	22	27.7
3	16.6	8	12.0	13	10.4	18	19.3	23	27.5
4	15.3	9	11.5	14	12.8	19	22.4	24	26.2

table 3. Kansas to Pakistan path parameters.

APR 15

	LAT	LONG		PATH SHORT	PATH LONG
BASE	37:00	100:00	DIST	6723	14877
			HEAD	11	191

TARGET NAME: PAKISTAN
TARGET PREFIX: AP

	SUNRISE	SUNSET	
BASE	12:08	01:13	GEO-MAG 8.0
TGT-AVE	01:03	13:55	10.7 cm 133.0
TGT-MIN	00:25	13:28	SUNSPOT 86
TGT-MAX	01:33	14:22	Q-FACT 4.9

MAXIMUM USABLE FREQUENCY

0	9.0	5	8.9	10	10.2	15	14.9	20	9.1
1	12.3	6	8.3	11	13.1	16	13.0	21	8.5
2	13.5	7	7.9	12	14.6	17	11.7	22	8.0
3	10.9	8	7.5	13	15.7	18	10.7	23	7.6
4	9.6	9	7.2	14	16.1	19	9.9	24	9.0

FREQUENCY OF OPTIMUM TRANSMISSION

0	7.7	5	7.6	10	8.6	15	12.6	20	7.8
1	10.5	6	7.1	11	11.1	16	11.1	21	7.3
2	11.4	7	6.7	12	12.4	17	9.9	22	6.8
3	9.3	8	6.4	13	13.4	18	9.1	23	6.5
4	8.1	9	6.1	14	13.7	19	8.4	24	7.7

flux is 63.75). The ability to perform a screen dump to get hard copy is also available, and that format will be used to answer some of the questions raised at the beginning of the article.

Kansas to Marshall Islands

Suppose you're a top Kansas contester who needs that KX6 prefix on 10 and 15 meters for the VK/ZL contest during the middle of October. You've determined that the geomagnetic *A* index is presently 8 and the 10.7 cm flux is 133. When do you look for that rare one? Once the coordinates, date, country, and the geophysical information have been entered, the computer will give the information shown in **table 2.**

The first thing to note is that the great circle distance is less than 6000 miles. Since this is within the 250 to 6000 mile range in which the MUF algorithm is most accurate, you can feel fairly comfortable about using the MUF/FOT data. By examining the MUF, you can see that the best time to start looking for a 15-meter path is approximately 1800 GMT, which corresponds to the KX6 sunrise. Looking for the KX6 on 10 meters, you should wait about two hours, according to the MUF. Because the FOT shows that 10 meters might not be too dependable, spending more time on 15 meters could prove more fruitful; you would probably look for KX6 on 15 first. If you got him, you'd then look for him on 10.

Kansas to Pakistan

Once you've received the greatest number of points ever accumulated in the VK/ZL contest, you decide that it would be good to talk to your friend in Pakistan. Assuming that the 10.7 cm flux and the *A* index have not changed, you have all the information you need to use the program. After inputting the needed data, the computer calculates the information shown in **table** 3.

Using **table** 3, the first thing you'll note is that the great circle distance is greater than 6000 miles, so you'll use the MUF and FOT data with caution. You'll see with some surprise that there is an overlap between your sunrise and sunset in Pakistan. Knowing that the best chance for a long path contact is when your sunset occurs slightly before sunrise in Pakistan, you see that you have a 20-minute window (01:13 - 01:33) for a possible try at that time. There are two peaks in the MUF; you're going to try them both to maximize your chances of hooking up with your friend. You'll also try 30 meters between 0100 and 0300 GMT and try 20 meters between 1200 and 1500 GMT.

Texas grayline determination

What grayline locations are available to a station in Texas during sunrise? Once you've entered your base coordinates and the date, the computer will provide the data listed in **table 4.**

table 4. Grayline countries available to Texas at sunrise on November 15. Base latitude is 030:00. The longitude is 097:00. Sunrise is at 12:52, and sunset is at 23:33. The grayline headings are 201 and 021. For a grayline width of 16.00 degrees, these are the "gray" countries.

COUNTRY	PREFIX	SS/SR	DISTANCE	HEADING
AFGANISTAN	YA	SS	14694	195
ALABAMA	W5-AL	SR	553	75
ÅLAND ISLAND	OH0	SS	17021	207
ARKANSAS	W5-AR	SR	383	37
ASIATIC R.S.F.S.R.	UA9-0	SS	16289	162
AZERBAIJAN	UD6	SS	15430	207
CASEY STATION	VK0	SS	13175	17
CHAGOS ISLANDS	VQ9	SS	12291	204
CLIPPERTON ISLAND	FO8	SR	1433	216
COLORADO	W0-CO	SR	699	323
DENMARK	OZ	SS	17179	214
DUMONT D'URAVILLE STATION	FB8Y	SS	13792	25
EASTER ISLAND	CE0A	SR	3492	193
ESTONIA	UR2	SS	16842	206
EUROPEAN R.S.F.S.R.	UA2	SS	15906	206
FINLAND	OH	SS	17060	203
FLORIDA	W4-FL	SR	744	96
GEORGIA	W4-GA	SR	688	72
GEORGIA (USSR)	UF6	SS	15666	209
GREENLAND	OX-XP	SR	3107	19
ICELAND	TF	SR	3495	29
ILLINOIS	W9-IL	SR	717	31
INDIA	VU	SS	13981	178
INDIANA	W9-IN	SR	797	38
IOWA	W0-IA	SR	732	10
IRAN	EP	SS	14930	206
JAN MAYEN	JX	SS	17912	202
KANSAS	W0-KS	SR	490	349
KARELO-FINNISH REPUBLIC	UN1	SS	16955	199
KAZAKH	UL7	SS	15500	191
KENTUCKY	W4-KY	SR	720	51
LABRADOR	VO2	SR	2077	36
LACCADIVE ISLANDS	VU7	SS	13262	194
LATVIA	UQ2	SS	16797	208
LENINGRADSKAYA STATION	UA1	SS	14253	24
LOUISIANA	W5-LA	SR	215	73
MALDIVE ISLANDS	8Q-VS9	SS	13035	197
MANITOBA	VE4	SR	1440	1
MARKET REEF	OJ0	SS	17034	206
McMURDO	KC4	SS	14369	14
MEXICO	XE	SR	604	208
MICHIGAN	W8-MI	SR	990	29
MINNESOTA	W0-MN	SR	915	9
MISSISSIPPI	W5-MS	SR	428	72
MISSOURI	W0-MO	SR	514	20
NEBRASKA	W0-NE	SR	671	350
NEW MEXICO	W5-NM	SR	517	300
NEW YORK	W2-NY	SR	1329	48
NORTH CAROLINA	W4-NC	SR	987	64
NORTH DAKOTA	W0-ND	SR	1033	352
NORTHWEST TERRITORIES	VE8	SR	2255	353
NORWAY	LA	SS	17379	209
OHIO	W8-OH	SR	911	45
OKLAHOMA	W5-OK	SR	243	351
OMAN	A4X	SS	14155	210
ONTARIO	VE3	SR	1273	23
PAKISTAN	AP	SS	14502	195
PENNSYLVANIA	W3-PA	SR	1154	50
QUEBEC	VE2	SR	1666	31
SCOTT STATION	ZL5	SS	14369	14
SOUTH DAKOTA	W0-SD	SR	856	350
SVALBARD	JW	SS	18119	194
SWEDEN	SK	SS	17166	207
TADZHIK	UJ8	SS	14938	190
TENNESSEE	W4-TN	SR	647	53
TEXAS	W5-TX	SR	184	312
TURKOMAN	UH8	SS	15180	201
UKRAINE	UB5	SS	16268	211
UNITED ARAB EMIRATES	A6X	SS	14393	211
UZBEK	UI8	SS	15364	203
VIRGINIA	W4-VA	SR	927	55
WHITE R.S.S.R.	UC2	SS	16546	210
WISCONSIN	W9-WI	SR	907	20

table 5. List of long path countries from Texas at sunrise on November 15. (These countries meet the additional criterion that Texas sunrise occurs after their sunset.)

AFGHANISTAN
EUROPEAN R.S.F.S.R.
FINLAND
IRAN
KARELO-FINNISH REPUBLIC
KAZAKH
NORWAY
PAKISTAN
SWEDEN
TURKOMAN
UZBEK

Assume that you're interested in working only long path this morning. If you know that the best chance for a long path QSO occurs between the distant station's sunset and your sunrise, use the program to determine which of the countries listed in **table 4** meet that condition. These countries are listed in **table 5.** The value given in the HEADING column is the long path direction to the particular country. The program gives long path when the sunrise or sunset of the distant QTH is opposite yours.

do it yourself?

This program is available for purchase in two versions (one with the grayline feature, DX1, and one without the grayline feature, DX2), each configured for any 64k Apple with a disk drive and 16K expander (a printer is optional; if you have one, be sure to plug it into slot No. 1). Purchase information, as well as information describing computer services available to owners of computers other than the Apple, is provided at the end of this article.

It took the two of us *nine months* to develop and debug this program. But if you'd like to try writing your own, here's what you'll need to know. Almost all the information any DX or SWL operator needs for effective operation is available in standard sources and can be written into a single program. You'll need a variety of resources — a thorough understanding of programming *your* computer — but it can be done.

Basic sources of data include a DX country list and for each country given, the appropriate coordinates; you'll also need a MUF summary, beam headings, Great Circle Distances, Sunrise and Sunset times, and a list of grayline countries. The grayline list can be generated using a variety of techniques, including calculation from sunrise/sunset times and longitude[3] or a slide rule-like product, *DX Edge*;* what makes this program unique is not the data itself, but how it is assembled and made useful.

developing the data base

From the beginning it was clear that the program would have to have a world data base that was always in memory. This would allow the other sections of the program using the data base to run at a higher speed, and eliminate the need for disk access, which would slow the program down. The program would also have to be done in a language that would allow the source code to be compiled; we chose Pascal, which allowed a structured approach to be followed and made "modular code development" possible. By using Pascal, we could write, debug, and compile each module separately. The program would also have to integrate the various data in such a way that would allow the program to be easy to use — i.e., "user friendly." It would have to "fit" into a 64k Apple.†

The most challenging part of designing this program was the development of the grayline feature; this is the area we'll discuss here in the greatest depth. The other sections of the program will be summarized, with the sources or the equations used to develop them shown in appendix A.

Using the equations shown in appendix A, you can write your own version of the program. *The version of the program shown in the examples in this article has been copyrighted*; however, all pertinent information is shown here for the benefit of *ham radio's* readers who wish to experiment with developing their own programs.

Our final world data base consists of 433 countries, states, and provinces, as well as prefixes and an average of three points per country for doing the grayline determination. A complete 15-page listing of the world data base will be provided for a 8½ × 11 SASE and $3.00 to help defray the cost of photocopying. (Send requests directly to author.)

The final program consists of 1600 lines of code or about 26 sheets of paper requiring 61,440 bytes of storage before compilation. After the data is compiled, an additional 21,504 bytes are needed for storage. Again, note that a large portion of this program is designed to produce "friendly" input and output routines; this results in a great deal of software "overhead."

The data base required 71,440 bytes of disk storage — more than our machine's memory could accommodate. By applying a "compaction scheme," we were able to reduce the number of bytes required to 19,456.

determining grayline

The grayline is the Great Circle path that defines or

*DX Edge is available from Ham Radio's Bookstore, Greenville, New Hampshire 03048 ($16.95 plus $1.50 postage and handling).

†A 16k extender card must be added for the version of the program reproduced here and the purchased program as well. The cost of a 16k card is approximately $25.00.

separates the regions of the world that are in daylight and darkness. Each location on the globe has two graylines each day; one at sunrise and the other at sunset. In order to accurately determine the grayline, an accurate equation for sunrise is necessary. Part of the sunrise/sunset equation is the declination of the sun; the equation that describes sunrise time and sunset time and declination of the sun are shown in appendix A.

In order to determine whether or not a particular country is on the grayline, a comparison must first be made between the sunrise/sunset of the desired location and the desired country. But which coordinates should be used to describe the location of a country? Consider, for example, the USSR. This country has several different radio prefixes within it and spans thousands of miles. It's obvious, then, that any "average" of the latitude and longitude would be grossly incorrect in describing the USSR. If the country selected were Andorra, then this means of describing the country would be fully acceptable. So what's the best way to describe each country? Each country must be described (i.e., digitized with a sufficient number of points) in such a manner that during the year all minimum and maximum sunrise/sunset points will have been obtained.

By trial and error, the following approach has been successful in limiting the total number of data points necessary to describe each country.

In the rough outline drawing of the state of California (**fig. 1**), the extreme points of the state are emphasized. These are the points that were assumed to be the minimum number necessary to describe California as far as differences in sunrise and sunset time are concerned.

All of the points on the outline of the state were programmed into the sunrise and sunset equations shown in the appendix along with intermediate points to confirm that they were, in fact, sufficient to describe a "radio country." This approach was determined to be valid. (While the simple drawing shown in **fig. 1** is far less complex than a sketch of the USSR — with its many prefixes — would be, for example, it should be easy to imagine what the final "radio country" sketch of such a vast area would look like.)

This approach was used for every country in the *DX Callbook*, for each one of the United States, for all provinces of Canada, and for any new country that could be found at the time the program was developed. Once the world data base was developed, a second data base was extracted from it and used to calculate beam headings and Great Circle distances. This second data base is essentially the average of all the given points that describe each country. In **fig. 1**, point E represents the average.

Readers who want to generate their own data base can produce the best result if they use a computer that automatically digitizes each country and then computes sunrise and sunset times throughout the year.

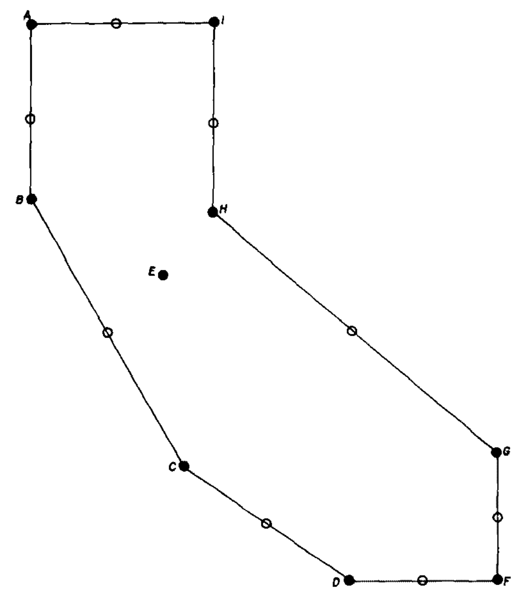

fig. 1. Calculation of sunrise and sunset times begins with a rough sketch of area to be identified.

using the data base

If the sunset or sunrise time is accurately known for your location, then it's easy to determine the countries that lie on the grayline. Calculate the sunrise and sunset times for each of the points of a particular country. Since the time that the grayline occurs is identical for all the points along it, compare each of the calculated times in the countries of interest to the time at your base location. If the difference in time is within the range of time desired — typically $\pm$ 1/2 hour, or if the base sunrise/sunset lies between the minimum and maximum rise/set times, then that particular country is on the grayline.

When printing out the grayline, the subroutines for heading and distance are also used as an aid in determining long and short path propagation. The beam heading for grayline can be determined by first using the equation of time rewritten to solve for longitude. The latitude used is arbitrarily chosen to be the target latitude + or − 25 degrees. Once the longitude is found, the equation for beam heading is employed by using your base location and the arbitrarily chosen latitude and computed longitude. This will then be the grayline heading. To output long path distance, just subtract the short-path distance computed by the equation in appendix A from 21600. The final item of

interest that can be extracted from the data base is *whether the country* that is on the grayline is going through sunrise or sunset. Keep track of which time equation was used to match your base location time — sunrise or sunset — and make note of which one it was when printing out your listing. A brief excerpt from the program manual will show why this information is important to you:

Daylight overlap. Compare your sunrise with the target sunset(s). The time span between could be good for daylight openings.

Daylight openings. Try 10 or 15 meters, except when the sunspot number is high.

Darkness overlap. Compare your sunset with the target sunrise(s). The time span could be good for darkness openings.

Darkness openings. Try 30, 40, 80, 160 meters.

Long-path openings. Base sunset occurs *before* target sunrise; base sunrise occurs *after* target sunset; use long-path heading.

Grayline openings. Eastern paths: look *before* sunrise; western paths: look *after* your sunrise; use grayline heading.

conclusion

With the writing of this program, we have attempted to combine a variety of aids for the Novice or experienced operator that will free up his or her schedule for more time on the air. For the first time, an accurate computerized grayline that leaves no doubt as to what country lies on the grayline and what time to operate is available. (This program will *not* help predict those cases in which what appears to be grayline propagation is occurring, but is one hour too early.) Propagation modes are varied and many; this program serves as a predictive tool to the ham operator and SWLer alike in providing assistance for some, but not all, modes of propagation.

acknowledgement

We would like to acknowledge the patience that both our wives exhibited while we developed and tried to perfect this program. Many thanks also to Vern, K6VXY, for his unflagging love of CW that gave us the idea and inspiration for this program.

references

1. Garth Stonehocker, KØRYW, "DX Forecaster," *ham radio*, August, 1982, page 80.
2. Robert Rose, K6GKU, "MINIMUF: A Simplified MUF-prediction Program for Microcomputers," *QST*, December, 1982, page 36.
3. D. Hoppe, K6UA, P. Dalton, W6NLZ, and F. Capossela, K6SSS, "The Grayline Method of DXing," *CQ*, September, 1975, page 27.
4. G.P. Stancey, G3MCK, "An Easy Method for Determining Grayline Bearings," *Radio Communication*, July, 1981, page 623.

appendix

Use the following equations to determine base and target sunrise/sunset times, long/short path output, MUF/FOT, and gray-line. Note that some equations include additional code; be sure to include these codes to prevent "crashing" when implementing this program on your computer.

Date-dependent constants. This little do-loop determines the day number of the present date: for example, January 1 is Day Number 1; December 31 is Day Number 365.

```
monum='303232332323'
FOR I  = 1 TO (MONTH - 1) DO
DAYNUM = DAYNUM + ORD(MONUM[I]) - 20;
DAYNUM = DAYNUM + DAY;
```

Sunrise/sunset constants.

```
pi = 3.14159265
pj = pi / 180
pk = 180 / pi
```

(*Lat* signifies latitude; *Lon* indicates longitude. Both are expressed in radians.)

```
DAYCONST = 0.017202 * (DAYNUM - 1.5)

EQ_OF_T = (-1.842 * SIN(DAYCONST-0.05952)
      -2.482 * SIN(2*DAYCONST+0.3557) - 0.079 * SIN(3*DAYCONST+0.2967)
      -0.055 * SIN(4*DAYCONST+0.6981) - 0.003 * SIN(5*DAYCONST+0.7156))
      / 15

DECL = PJ * (0.379 - 23.267 * COS(DAYCONST+0.1793)
      -0.381 * COS(2*DAYCONST+0.1292) - 0.171 * COS(3*DAYCONST+0.5184)
      -0.008 * COS(4*DAYCONST+0.4538) - 0.003 * COS(5*DAYCONST+1.658))

temp = 12 - eq_of_t + long * pk / 15
temp1 = arccos((-0.01454 - sin(lat) * sin(decl)) /
   (cos(lat) * cos(decl))) * pk / 15;

sunrise = temp - temp1
sunset = temp + temp1
```

Answer is in decimal hours.

Conversion of 10.7 cm flux and the quality factor.

```
sunspot number = (-0.728 + sqrt(0.728*0.728 - 4 * (63.75 - flux) * 0.00089))
/ 0.00178;

qual   := 1.0857 * ln(flux) * (1.0 - 0.2625 * cos(8.642E-3 * daynum) *
cos(8.642E-3 * daynum)) * exp(-0.01 * gflux) + 0.82;
```

g **flux** indicates geomagnetic flux or A factor; **flux** signifies the 10.7 cm flux as reported by WWV.

Great Circle distance/beam headings. (Distance expressed in radians*; heading from Lat 1, Lon 1 to Lat 2, Lon 2. Check for poles.)

```
IF LAT1 > 1.5533 THEN LAT1 = 1.5533
IF LAT1 < -1.5533 THEN LAT1 = -1.5533

IF LAT2 > 1.5533 THEN LAT2 = 1.5533
IF LAT2 < -1.5533 THEN LAT2 = -1.5533
```

*To convert distance to nautical miles, multiply by 60 * pk.

Check for diametrically opposed locations.

```
IF (ABS(LON1 - LON2) > 3.124139) AND
((-1.0 * ROUND(LAT1*120)) = (ROUND(LAT2 *120)))

THEN LON2 = LON2 + 0.01745
```

Check for equal longitudes.

```
IF ROUND(LON1 * 120) = ROUND(LON2 * 120) THEN DISTANCE = ABS(LAT1 - LAT2)
IF LAT1 > LAT2 THEN HEADING = PI ELSE HEADING = 0.0


DISTANCE = ARCCOS(SIN(LAT1) * SIN(LAT2) +
           COS(LAT1) * COS(LAT2) * COS(LON2 - LON1))

HEADING  = ARCCOS((SIN(LAT2) - SIN(LAT1) * COS(DISTANCE)) /
            (SIN(DISTANCE) * COS(LAT1)))
```

Check to see if heading needs to be reversed.

```
IF SIN(LON2 - LON1) > 0.0 THEN HEADING = 2 * PI - HEADING
```

For information on how to order a copy of the program, or to obtain other DX computer services if you don't own an Apple, contact Van Brollini, NS6N, 5861 Bridle Way, San Jose, California 95123 (SASE appreciated).

ham radio

✓ 138

ham radio TECHNIQUES

Bill Orr W6SAI

As of this writing (mid-June) it looks as if the FCC is slowly moving toward the possible release of additional frequencies for Amateur Radio, as authorized by the 1979 World Administrative Radio Conference. A good guess is that the so-called 24-MHz band may be released on Special Temporary Authority in the near future.

Interest in the new bands has received a boost because of the ARRL Antenna Design Competition, announced this past spring.[1] The goal of the contest is to provide new antenna designs that can be constructed at home and will cover as many of the WARC-79 bands as possible. The two entry categories include a five-band design that will cover all Amateur bands between 14 and 30 MHz and a six-band design that adds the 10-MHz band. (Details are listed in **table 1**; a "design frequency" indicating the midpoint of each band is also listed.)

table 1. Most antennas built for the design frequency specified will work well over the whole Amateur band. Two design frequencies are chosen for 10 meters, one for operation at the low frequency end of the band and one for operation at the high frequency end.

band (MHz)	design frequency (MHz)
10.100-10.150	10.12
14.000-14.350	14.17
18.068-18.168	18.11
21.000-21.450	21.22
24.890-24.990	24.94
28.000-29.700	28.60 (low)
	29.20 (high)

the "multiband" antenna

The problem of designing a single antenna that covers the frequency range of 14 to 30 MHz (or 10 to 30 MHz, as the case may be) is an interesting one. No doubt many fascinating antenna designs will surface during the next few months. One basic choice that must be made is whether the antenna designed will be a wideband type that covers the entire frequency span, or a type that functions only over the bands in question. Let's look at the wideband concept first.

For this discussion, a wideband antenna is defined as one that has relatively constant gain, polarization, and

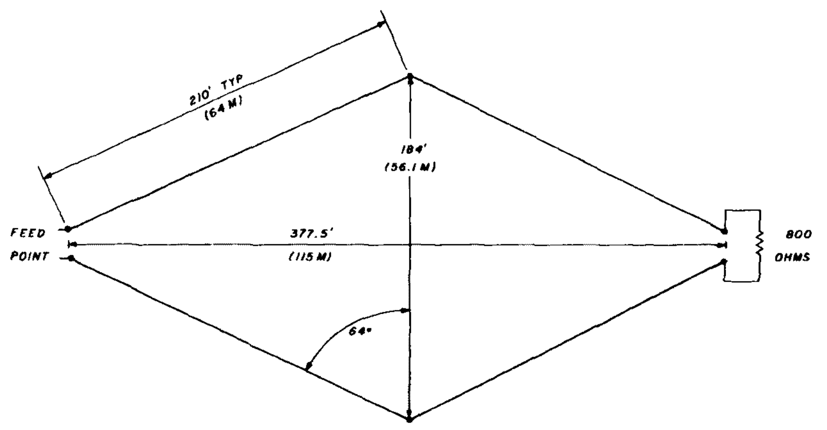

fig. 1. A terminated rhombic antenna for the frequency range of 14 to 30 MHz. Antenna is terminated with a noninductive resistor capable of dissipating about half the transmitter average power output. (Design taken from *The ARRL Antenna Book*, 14th Edition, Chapter 7, fig. 16.)

mpedance characteristics over the ɔperating range.

One of the earliest wideband gain ɜntennas is the terminated rhombic[2] ;hown in **fig. 1**. This antenna works ᴧell over a 2-to-1 frequency range, ɔrovides good front-to-back ratio and ;an exhibit as high as 14 dB gain over ɜ dipole, providing the leg-length of :he array is long enough. The front-to-ɔack ratio is achieved by choice of leg ength and the use of a terminating re-;istor, at the far end of the array, ᴧhich absorbs reflected power. Be-;ause the feedpoint resistance is about '50 to 800 ohms, an impedance match-ng transformer is required to match he antenna illustrated in **fig. 1** to a ;0-ohm transmission line.

Other varieties of traveling-wave an-ennas, such as the terminated V-an-enna, exist, but their main disadvan-age is their large size and the corre-;ponding amount of real estate they equire.

he log-periodic array

Two relatively new types of wide-ɔand antennas, the equiangular anten-ıa and the log-periodic antenna, are nore practical for Amateur service.

The equiangular antenna evolved 'om the observation that the proper-es of an antenna (impedance and pat-ɜrn) are determined by its shape and imensions with respect to operating ᴧavelength. When the antenna is caled, its properties are independent f frequency, provided its form is spec-ied only by angles and not by any par-cular dimension. A two-arm equian-ular spiral antenna, shown in **fig. 2**, ; an example of this design.

This antenna is a variation of the di-ole, where the two halves have been ᴧisted into a pair of equiangular arms. he antenna is fed by a balanced line t the center point. High frequency utoff is determined by antenna con-guration near the feedpoint; low fre-uency cutoff is determined by the uter circumference of the spiral, in-icated by the dashed line. Feedpoint ;sistance of the antenna is quite high, ɜing of the order of 120 to 180 ohms, ɜpending upon antenna size and de-

fig. 2. The two-arm *equiangular* dipole antenna is fed with a balanced line at F-F. The feedpoint impedance is relatively independent of frequency over operating range. Low frequency operating limit is defined by the circumference of antenna, indicated by the dashed line. (Drawing adapted from Jasik[2].)

sign. Frequency spans as great as 10-to-1 have been achieved in practice with this antenna type.

A second wideband antenna design, with which most Amateurs are familiar, is the log-periodic antenna (**fig. 3**).* The geometry of this antenna design is chosen so that the electrical properties repeat periodically with the logarithm of frequency. The basic trapezoidal-tooth log periodic structure is shown in the illustration.

The design can be further simplified if the structure is replaced by dipole elements (**fig. 4**). This log-periodic dipole antenna is a popular design for TV and FM receiving antennas, and versions of this antenna are often used by Amateurs on the VHF bands. The frequency limits of this antenna are those at which the outer elements are about one-half wavelength long.

The dipoles are fed at the center from a parallel wire transmission line in such fashion that successive dipoles have 180-degree phase reversal between them. A broadband structure is formed, with most of the radiation coming from the section containing elements approximately half a wavelength long at the operating frequency. Gain and bandwidth bear a definite relationship to the length and included angle of the antenna.

Unfortunately, at any given frequency in the operating passband, some of the elements in this array are inactive; the active element area moves along the structure as the frequency of operation is changed. At the lowest operating frequency, the longest elements have the most current in them and, as operating frequency is raised, the center elements become active and have the greatest current in them. At the upper frequency of operation, the shortest elements have the greatest current in them, with the longer elements relatively inactive. Thus a log-periodic dipole beam antenna must be considerably longer than a parasitic Yagi antenna of equivalent gain.

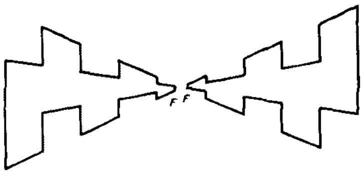

fig. 3. The geometry of the planar, log-periodic dipole antenna repeats periodically with the logarithm of the frequency.

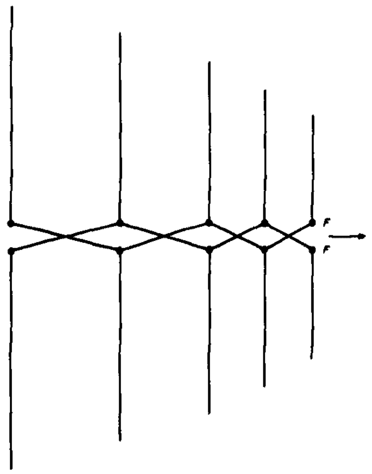

fig. 4. The log-periodic dipole array, in which linear elements replace the "teeth" of the antenna shown in *fig. 3*. This structure forms a beam antenna.

*Designed by KLM Electronics, Morgan Hill, California 95037.

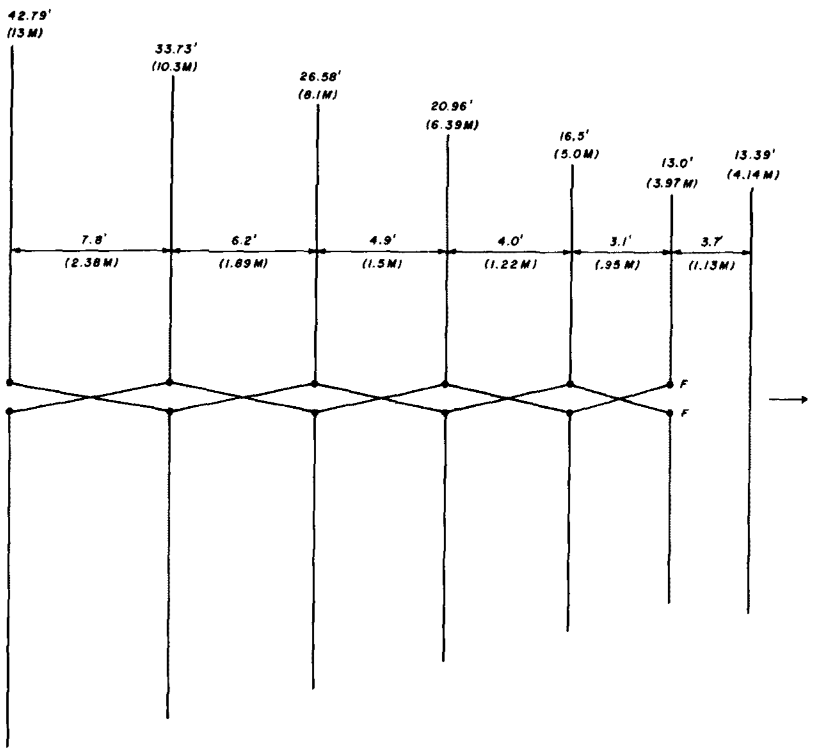

fig. 5. the KLM 13-30 LPY array covers 13-30 MHz with an average power gain of 5 dB over a dipole. The average front-to-back ratio is 15 dB. The array uses a single parasitic element.

the log-periodic Yagi

An interesting version of the log-periodic design is the log-periodic Yagi (LPY) antenna (**fig. 5**). This passband array provides higher gain per unit of length than the original designs, making use of a log-periodic dipole structure having the frequency characteristics of a bandpass filter. One or more parasitic elements are used to boost the gain of the log-periodic structure. Trimming the parasitics to specific frequencies can enhance the gain of the array at these frequencies at the expense of gain loss at other frequencies. Thus, an LPY antenna can be designed for maximum performance in closely adjacent Amateur bands (14, 18, 21, and 24 MHz, for example).

Other interesting wideband antennas exist but are not practical for use in the HF spectrum because of their size.

narrowband tuned antennas

The easiest way to get on one of the proposed new HF Amateur bands is to use a dipole cut to the midpoint of the band. A more useful antenna is a half-wave wire cut to the lowest band (for example, 10 MHz) and fed at the center with an open wire transmission line and an antenna tuner (**fig. 6**).

This is a very simple antenna and, when properly built and tuned, will cover all Amateur bands between 10 and 29.7 MHz. In fact, if the tuner is flexible enough, this antenna can also be used for the 40 and 80-meter bands, thus making a true "all-band" HF antenna.

At my station, I have an old Johnson Kilowatt "Matchbox" tuner that I picked up at a flea market. No longer made, this useful device will match almost anything at any frequency in the HF region. With this, or an equivalent tuner, the center-fed antenna dimen sions are not critical at all because th tuner makes up for variation in anten na and feeder length. If difficulty i experienced in loading up a particula combination of flat-top/feede lengths, adding or subtracting a foo or two of feeder length will usuall cure the problem.

parallel-connected dipoles

A simple multiband antenna used i the HF region consists of two or mor dipoles connected in parallel at th feedpoint. The ends of the dipoles ar separated a foot or two. This arrange ment works well when the bands ar harmonically related (7, 14, and 2 MHz, for example), but problems aris when the principle is applied to th new HF bands. Dipoles for 18 and 2 MHz, for example, when paralleled i this fashion do not seem to perforı properly. The 18-MHz dipole is una fected, but the 24-MHz dipole is con pletely detuned and will not perform a all! Other combinations have not bee tried, to my knowledge, but perhap one of *ham radio*'s readers will try di ferent combinations, such as 10 and ' MHz, or 14 and 24 MHz. I think th parallel-dipole approach has merit, b I haven't hit upon the lucky combin tion that works for the new bands

the multiband loop antenn

One interesting antenna that w cover five or six bands when used wi an open wire line and tuner is tl quarter-wave loop antenna (**fig. 7** With each side cut to 24 feet 6 inch (7.47 meters), the loop will cover bands from 10 MHz up through meters. Most loops of this type a mounted in the vertical plane and f at the bottom to provide horizontal p larization. Some experimenters ha had luck with this loop mounted in t horizontal plane, about 30 to 40 fe above the ground. Horizontal polariz tion is still provided.

If operation is desired only on MHz and up, the sides of the loop c be reduced to 13 feet 9 inches (4 meters).

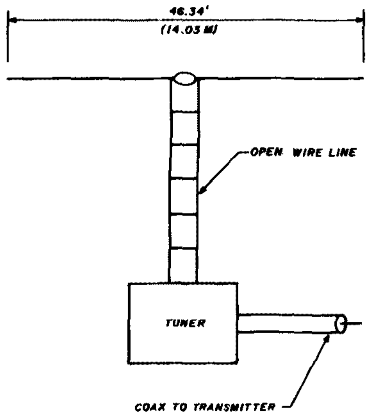

fig. 6. The 10 MHz dipole fed with random-length open wire line provides operation on all bands between 10 and 30 MHz if proper tuning unit is used.

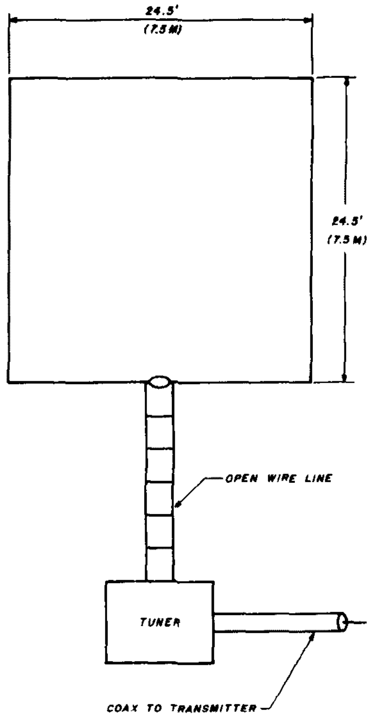

fig. 7. A simple quad-type loop antenna that will cover all bands from 30 through 6 meters. Open-wire line and antenna tuner provide balanced feed from a coaxial system.

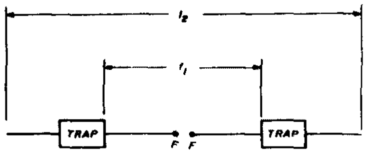

fig. 8. Trap dipole employs electrical "switches" to connect and disconnect antenna tips automatically for operation at frequencies f1 and f2.

the trap dipole

The trap dipole (**fig. 8**) makes use of the high impedance of a parallel resonant circuit to isolate or decouple unwanted tip portions of the antenna. The inner set of traps is placed in the element to isolate the center portion for operation on the highest frequency band (f_1). A second set of traps may be placed somewhat farther out along the element to isolate a longer portion, with the first set of traps becoming part of the antenna element at the lower operating frequency (f_2). Trap dipole antennas have been built with eight traps to allow operation on four Amateur bands. Trap design is straightforward, but determining the length of the tip sections, and the wire length between the traps is usually done on a cut-and-try basis.*

An approximate system for mathematically determining the length of the tip sections has been described in the Amateur literature.[3]

trap construction

Traps can be built either with discrete components (inductors and capacitors) or by using a length of coaxial line as a combined inductor and capacitor. Gary O'Neil, N3GO, described an interesting coaxial cable trap in *ham radio*, just a few years ago.[4] While his design provided somewhat better operational bandwidth than the conventional trap design, unfortunately, any form of trap made of coaxial cable is very difficult to accurately adjust to frequency because any variation in the spacing of the coiled cable can change the resonant frequency of the trap many hundreds of kilohertz. Trap construction and adjustment become quite a problem.

The trap made up of a capacitor and a separate inductor is easier to adjust to frequency, which is usually chosen as the midpoint of the band. A fixed, high voltage capacitor is commonly used; one popular unit is a 25 pF, 5 kV ceramic type.* Frequency can be adjusted by pruning the parallel-connected coil. Many Amateurs use prewound, spaced air inductors mounted on four plastic strips. One type of coil, the "miniductor" manufactured by Barker & Williamson Co., is suitable for this purpose.†

Unfortunately a trap made up of an air coil and a ceramic capacitor must be protected from the weather. Water can damage the capacitor, and ultraviolet light from the sun can quickly ruin the plastic strips supporting the coil! Solving these problems is not an easy task, and any ideas supplied by the readers as to the design of a weatherproof trap assembly would be appreciated.

practical two-band dipole for 18 and 24 MHz

Here's a simple antenna you can build in anticipation of the happy day when the 18 and 24-MHz bands are made available to Radio Amateurs for general communication. Important antenna dimensions are shown in **fig. 9**. The traps are made of a coil-capacitor combination, as discussed previously, and mounted to a small ceramic insulator which serves to take the pull of the antenna.

Before the traps are installed, they must be frequency-checked with a dip oscillator and a calibrated receiver. Place the trap in an area free of nearby metallic objects and loosely couple the dip oscillator to it. When reso-

*An article on a computer-aided design for a trap antenna is scheduled for publication in October. — Editor

*The popular Centralab type 850 capacitors are no longer made by this company. I understand an equivalent type is manufactured by Jennings Radio Co., 970 McLaughlin Avenue, San Jose, California 95122, and also by High Energy Corp., Lower Valley Road, Parkesburgh, Pennsylvania 19365.

†Barker & Williamson Co., 10 Canal Street, Bristol, Pennsylvania 19007

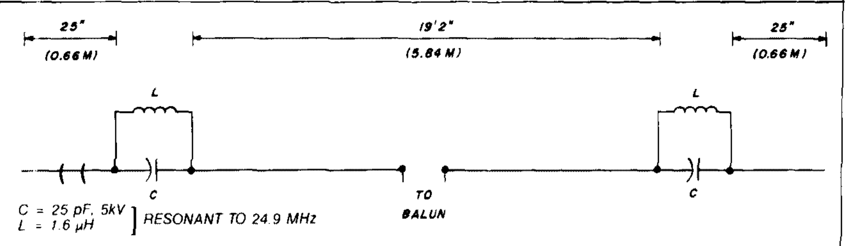

fig. 9. Trap dipole antenna for the 18 and 24 MHz bands. The trap is held in position by a 4-inch long ceramic insulator. The antenna is fed from a 1-to-1 balun and 50-ohm coaxial line.

nance is indicated, note the frequency of the oscillator on the nearby receiver. The traps should show resonance within ± 100 kHz of 24.9 MHz. One end turn on each trap should be broken free of the coil bars so that it can be moved about to set the exact resonant point of the trap. You'll find that when you attach the trap across the supporting insulator, the resonant frequency will drop a bit because of the capacitance across the insulator. It's best to shoot for a resonant trap frequency about 24.9 MHz; the insulator capacitance will then place your trap "right on the nose." You can also run your checks after assembly — take your choice.

The length of the other tip sections is critical for proper operation on 18 MHz. Varying the tip length by as little as one inch per end will change the resonant frequency about 150 kHz. Since the band is only 100 kHz wide, this means tip dimensions are critical to about an inch to establish antenna resonance within the band.

The tip dimensions shown in **fig. 1** are quite accurate. If you want to frequency-check the whole antenna, suspend it in the air, in the clear, about six feet above the ground. Place a 1/2-turn inductor across the center insulator, and measure the 18 and 24 MHz frequencies of the complete antenna with a dip oscillator coupled to the inductor.

When I made my antenna I cut the tip sections about a foot longer than necessary and then folded them back and twisted the wires around the active antenna wire. That provided plenty of extra wire in case I had to lengthen the tip sections. Once I reached the correct length, I cut off the excess wire.

After some minor adjustments were made I found out that removing one inch at each end of the center dipole raised the resonant frequency of the antenna 100 kHz/inch at 24.9 MHz. (Since the length of the 24.9 MHz dipole affects the resonant frequency of the 18.1 MHz dipole, pruning the 24.9 MHz dipole must be done before the tip sections are adjusted.)

When the antenna was completely tuned, it was hauled up my tower and anchored at the 45-foot level, with the ends dropping down to the 25-foot level and tied to two nearby trees. SWR measurements revealed that the maximum figure on either range was under 1.3-to-1, with near-unity SWR at the design frequency on each band.

Note: More information on multiband antennas and trap antennas can be found in the 22nd edition of ***Radio Handbook****, available from Ham Radio's Bookstore, Greenville, NH 03048, at $21.95, postpaid.*

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ham radio

applied Yagi antenna design part 4: a 50-MHz Tilton/Greenblum design

Computer model analyzes, updates an atypical Yagi design

Of the four VHF/UHF bands discussed in this series, 50 MHz is the most difficult mechanically. Where the other bands' Yagis are able to employ welding rods for the driven and parasitic elements, 50 MHz requires tubing of at least 0.5 inches (1.27 cm) in diameter. Fortunately, long enough single lengths of tubing are available so the builder can avoid the additional calculations needed to compensate for telescopic elements.[1] While element mounting may require methods that have some effect on element lengths, these effects are known quantities, and of relatively minor consequence.[2]

Because the NBS 1.2 wavelength design is often the practical limit, long-boom Yagis are not common on 50 MHz. Larger Yagis would be difficult to impossible for most 50 MHz operators; stacking would present even *more* difficulties. In effect, 50 MHz presents the VHF operator many of the antenna problems normally associated with the HF bands.

selecting a 50 MHz design

Lawson demonstrated that the simple Yagi performed almost as well as any other antenna in the lengths feasible on 50 MHz.[3] Any significant improvement would have to be derived from an unequal spacing approach for the parasitic elements. Tilton presented one such design, and since its initial publication 27 years ago, it has become a standard of comparison, even being adopted by a commercial antenna manufacturer.[4] As is true of Tilton's other fine antennas, his six-element 50 MHz classic is based on the Greenblum design data.

Simply adding another director to Tilton's antenna for iteration purposes would result in a higher calculated gain and hopefully in a better overall pattern. But as is evident from examining Greenblum's data, there is a difference between an optimized seven-element Yagi and a six-element Yagi with an added element.

Tilton experimented with Greenblum Yagis of up to eleven elements. Of these Yagis, a seven-element 220 MHz design was published.[5] The design frequency appears to be 221.5 MHz, and a non-conductive boom is used. Scaling to the 50 MHz band is easily accomplished, permitting the iteration process to begin. A design frequency of 50.25 MHz was selected since a Yagi centered on this frequency can easily provide good performance across the weak signal area. **Table 1** presents spacing data for this Yagi; column two of that table contains the inter-element spacings that were calculated from those given for the 220-MHz design. The former Yagi's spacings were stated in inches and were converted to wavelengths at 221.5 MHz. These wavelengths were converted to the comparable number of inches at 50.25 MHz. For the builder's convenience, the 50.25 MHz spacings were rounded to the nearest 0.125 inch. Column three of **table 1** shows the cumulative addition of the spacings in column two, and column four is the restatement of column three in wavelengths at 50.25 MHz. All element diameters are 0.5 inches, as used in Tilton's classic. Some builders may want to use 0.75-inch tubing as is done in *The Radio Amateur's Handbook* for the NBS Yagi of this same boom length.[6]

iterating the 50-MHz Yagis

Reflector and director lengths were incremented in

table 1. The baselined 50.25 MHz Yagi with parasitic element lengths supplied during iteration, and the driven element fixed at the non-reactive length of 112.545495 inches.

element name	element spacing (inches)	cumulative spacing (inches)	cumulative spacing (wavelengths)
reflector	0.000	0.000	0.0000
driven	44.000	44.000	0.1873
director 1	33.000	77.000	0.3278
director 2	37.500	114.500	0.4875
director 3	45.500	163.000	0.6940
director 4	57.250	220.250	0.9377
director 5	66.125	286.375	1.2192

By **Stanley Jaffin, WB3BGU**, 800 Stonington Road, Silver Spring, Maryland 20902

table 2. A comparison of optimized 50.25 MHz Yagis for each of six different director tapering approaches.

taper	optimized performance parameter	parasitic element length in inches reflector	director 1	gain (dBi)	F/B (dB)
0.000	gain	116.000	105.750	12.493	15.192
	F/B	115.500	103.000	12.066	49.580
0.125	gain	116.000	106.000	12.481	15.624
	F/B	115.750	103.250	12.066	50.774
0.250	gain	116.000	106.500	12.469	15.109
	F/B	115.750	103.500	12.063	59.342
0.500	gain	116.000	107.000	12.441	15.843
	F/B	116.000	104.000	12.057	52.097
1.000	gain	116.000	107.750	12.375	17.878
	F/B	117.000	104.250	11.884	60.431
1.500	gain	116.250	108.500	12.301	19.108
	F/B	117.500	104.750	11.769	42.259

table 3. Optimized gain iteration for a taper of 0.000 with a reflector length of 116.00 inches.

director 1 (inches)	gain (dBi)	F/B (dB)
102.00	11.859	30.072
102.25	11.193	32.852
102.50	11.967	36.683
102.75	12.020	39.984
103.00	12.074	36.812
103.25	12.127	32.401
103.50	12.179	29.056
103.75	12.229	26.464
104.00	12.278	24.352
104.25	12.325	22.564
104.50	12.366	21.007
104.75	12.405	19.623
105.00	12.438	18.372
105.25	12.464	17.228
105.50	12.483	16.173
105.75	12.493	15.192
106.00	12.492	14.276
106.25	12.479	13.417
106.50	12.453	12.610
106.75	12.411	11.852
107.00	12.352	11.140
107.25	12.276	10.473
107.50	12.183	9.852
107.75	12.072	9.280
108.00	11.944	8.759
108.25	11.802	8.296
108.50	11.648	7.899
108.75	11.486	7.580
109.00	11.320	7.358
109.25	11.154	7.261
109.50	10.989	7.332
109.75	10.822	7.646
110.00	10.633	8.340

table 4. Optimized F/B iteration for a taper of 0.000 with a reflector length of 115.50 inches.

director 1 (inches)	gain (dBi)	F/B (dB)
102.00	11.856	27.029
102.25	11.909	29.117
102.50	11.961	32.046
102.75	12.014	36.815
103.00	12.066	49.580
103.25	12.118	41.466
103.50	12.169	33.637
103.75	12.218	29.414
104.00	12.266	26.457
104.25	12.312	24.152
104.50	12.354	22.247
104.75	12.393	20.612
105.00	12.427	19.172
105.25	12.454	17.881
105.50	12.475	16.708
105.75	12.486	15.631
106.00	12.488	14.634
106.25	12.478	13.707
106.50	12.454	12.841
106.75	12.416	12.032
107.00	12.362	11.275
107.25	12.290	10.570
107.50	12.201	9.915
107.75	12.093	9.312
108.00	11.969	8.764
108.25	11.829	8.277
108.50	11.676	7.859
108.75	11.513	7.522
109.00	11.344	7.285
109.25	11.172	7.175
109.50	10.998	7.236
109.75	10.818	7.542
110.00	10.613	8.231

0.25 inch steps, and element tapering was initially incremented in 0.5 inch steps. As a result of obtaining what appeared to be strange results for Greenblum Yagis, iterations were also made for tapers of 0.125 and 0.25 inch. All of these results are summarized in **table 2**. A zero taper clearly gives the highest calculated gain figures, even though Greenblum Yagis usually require some degree of parasitic element

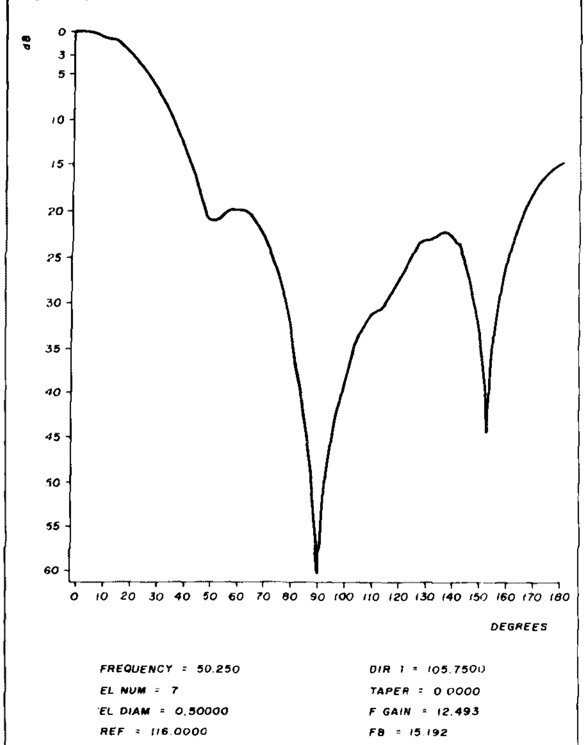

fig. 1. Zero taper gain optimized E-plane plot.

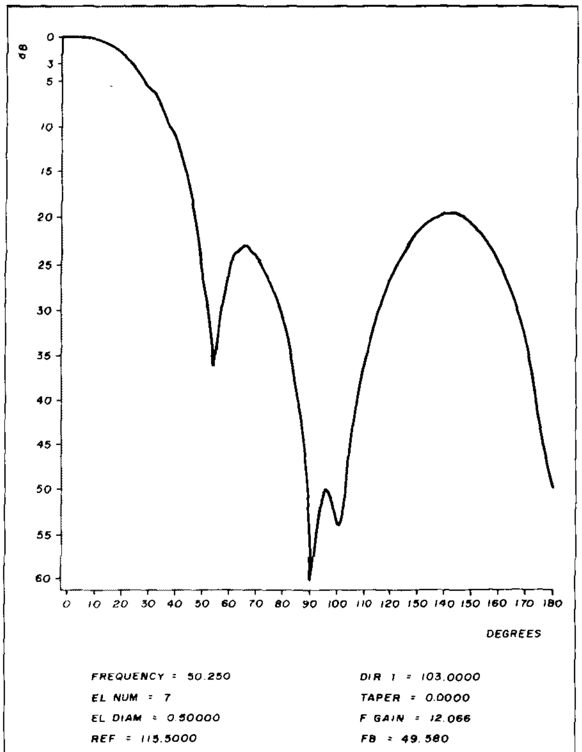

fig. 2. Zero taper F/B optimized E-plane plot.

table 5. Frequency response parameters for the gain optimized Yagi with a 0.000 taper.

frequency (MHz)	gain (dBi)	F/B (dB)
47.25	9.477	7.003
47.75	10.299	9.345
48.25	10.987	12.406
48.75	11.542	16.781
49.25	11.993	23.899
49.75	12.337	22.019
50.25	12.493	15.192
50.75	12.314	11.005
51.25	11.746	8.321
51.75	11.083	7.181
52.25	10.671	9.386
52.75	3.191	−0.549
53.25	−8.390	−14.057

table 6. Frequency response parameters for the F/B optimized Yagi with a 0.000 taper.

frequency (MHz)	gain (dBi)	F/B (dB)
47.25	8.775	5.016
47.75	9.672	6.858
48.25	10.431	9.105
48.75	11.014	11.874
49.25	11.449	15.540
49.75	11.787	21.460
50.25	12.066	49.580
50.75	12.282	21.748
51.25	12.378	15.236
51.75	12.247	11.331
52.25	11.816	8.631
52.75	11.204	6.984
53.25	10.835	7.071

tapering.[7] The zero taper F/B optimized Yagi, while not possessing the maximum calculated F/B, does have an outstanding F/B that is clearly representative of the F/B optimized Yagis.

Table 3 presents the iterations that identified the optimized gain of 12.493 dBi. **Table 4** presents the iterations that identified the optimized F/B of 49.580 dB. **Tables 5** and **6** present these antenna's respective frequency performance parameters, in 500 kHz increments across a 6 MHz bandwidth.The gain optimized antenna shows a clear peak at 50.25 MHz. The F/B optimized antenna also shows a peak in the optimized parameter at 50.25 MHz, almost to the point of being of the single frequency vectorial cancellation type of F/B. **Figs. 1** and **2** display these antennas' respective E-plane plots.

The comparison between these two Yagis serves to emphasize the superiority of the F/B antenna. This is the Yagi with the obviously more clearly defined main lobe and the reduced-amplitude minor lobes. With the

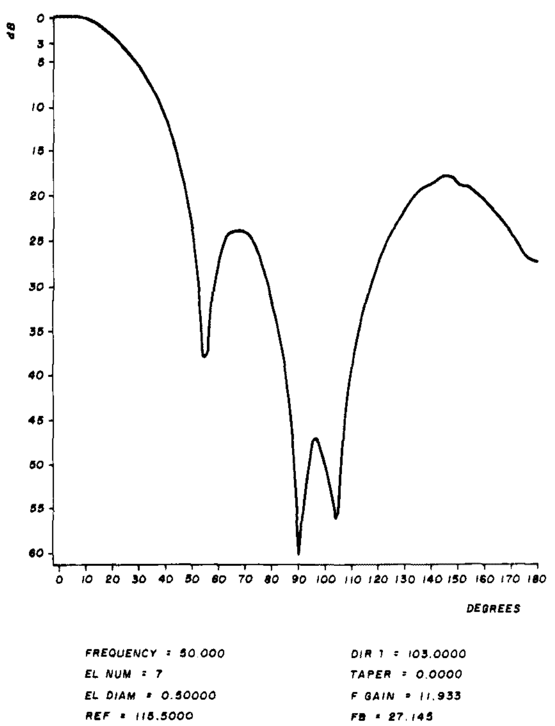

fig. 3. F/B optimized Yagi at 50.00 MHz.

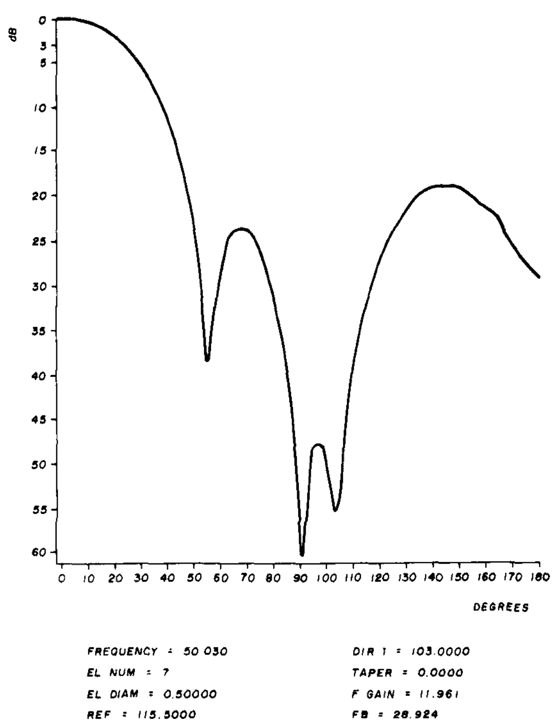

fig. 4. F/B optimized Yagi at 50.05 MHz.

exceptions of a slightly narrower main lobe and additional unwanted signal attenuation between 132 and 158 degrees, the gain optimized Yagi is a second choice. While unwanted signals are rarely an exact 180 degrees away from desired signals, a 15 dB F/B is more or less marginal for a Yagi of this length. However, the adequacy of this F/B ratio is really a function of band activity and operator preference.

A final question, addressing the bandwidth over which the F/B optimized antenna's F/B ratio and general pattern can be realized, remains. **Table 7** presents this Yagi's performance parameters as calculated from 50.0 MHz to 50.5 MHz, in 50 kHz increments. **Figs. 3** through **13** contain the E-plane plots that correspond to each 50 kHz increment, with that for 50.25 MHz being repeated for purposes of clarity. Over the frequency range of interest to weak signal operators, calculated F/B begins at 31.168 dB, peaks at 49.580 dB, and drops to 33.170 dB. Calculated gain begins at 11.988 dBi and rises to 12.139 dBi. These figures serve a second purpose, as they provide a study in the effects of slight variances in frequency response on the performance criteria on which a given Yagi was optimized. It is interesting to note the emergence of another minor lobe starting at 50.3 MHz.

A comparison with the NBS 1.2 wavelength Yagi proves the value of either of the optimized Tilton/Greenblum designs. While only 0.0192 wavelengths

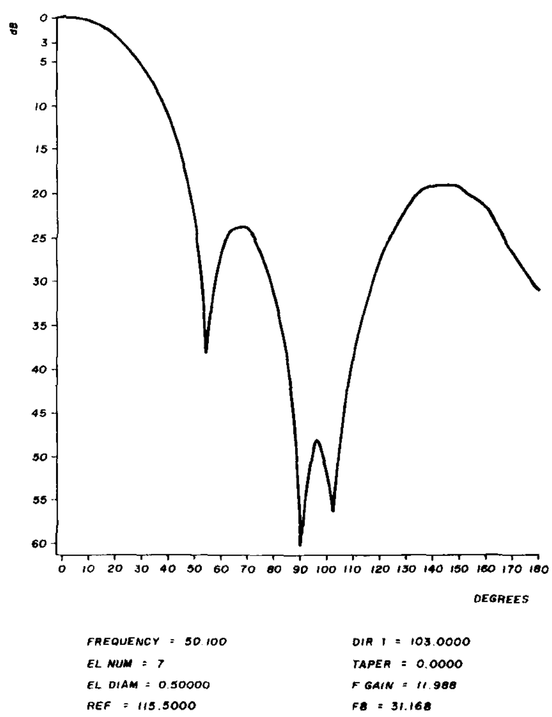

fig. 5. F/B optimized Yagi at 50.10 MHz.

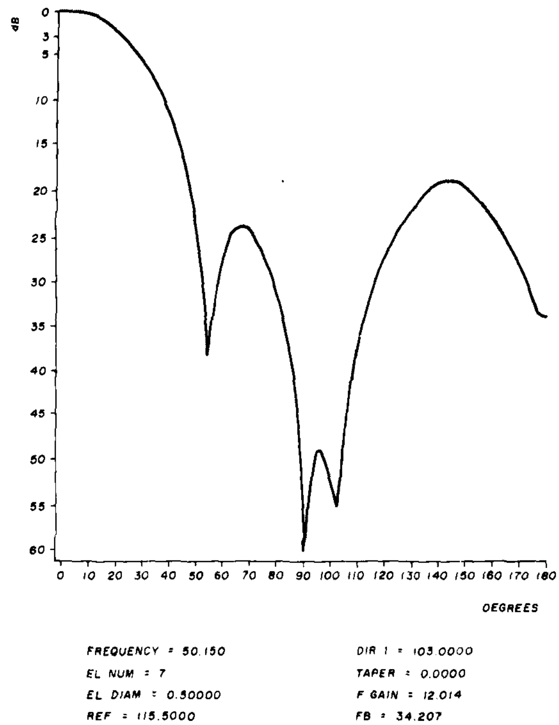

fig. 6. F/B optimized Yagi at 50.15 MHz.

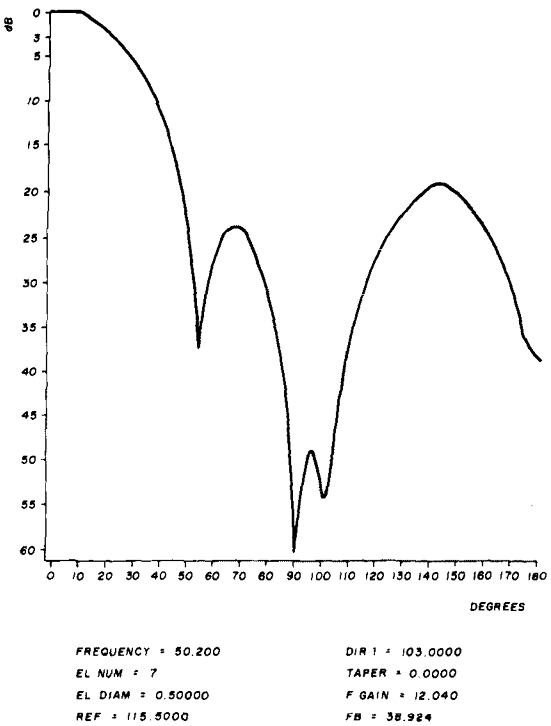

fig. 7. F/B optimized Yagi at 50.20 MHz.

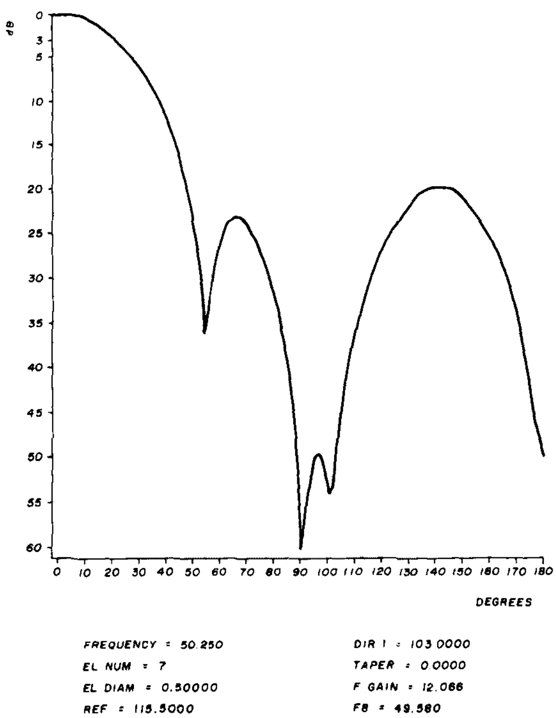

fig. 8. F/B optimized Yagi at 50.25 MHz.

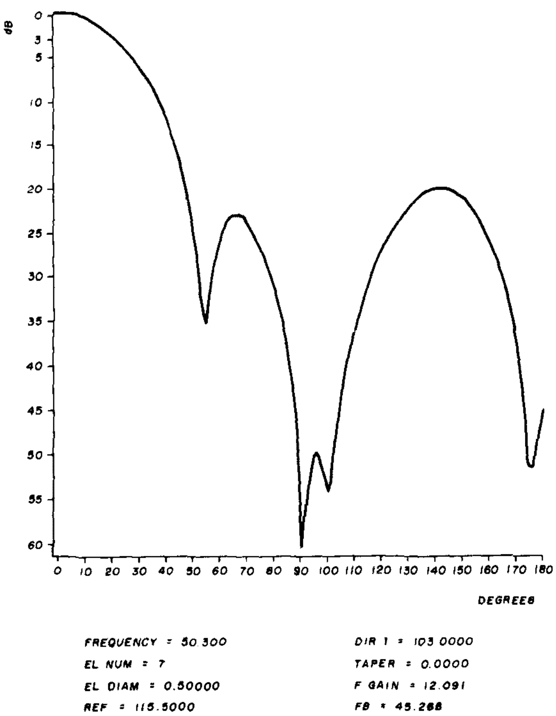

fig. 9. F/B optimized Yagi at 50.30 MHz.

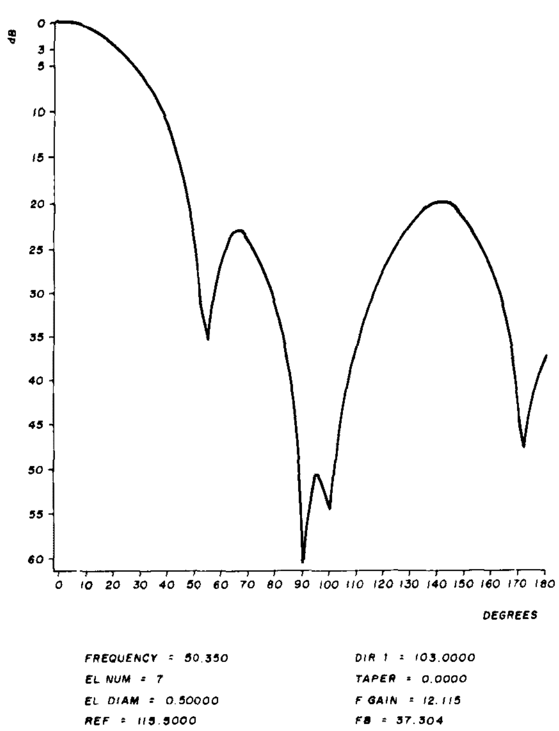

fig. 10. F/B optimized Yagi at 50.35 MHz.

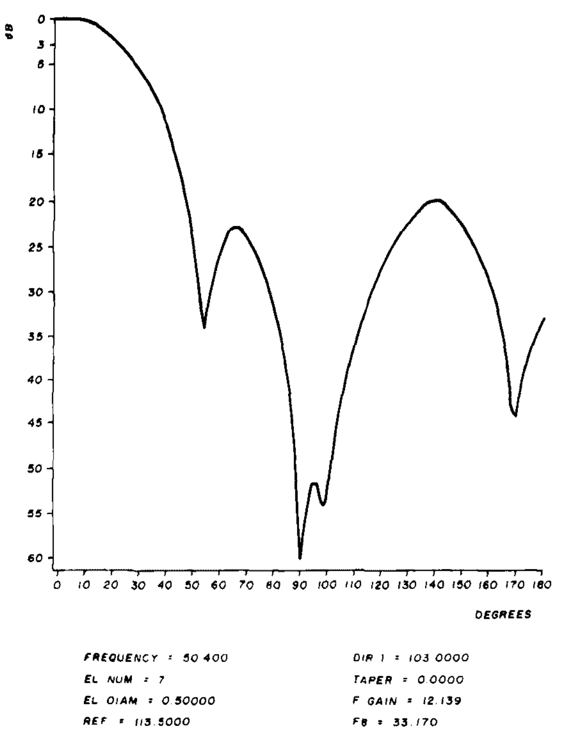

fig. 11. F/B optimized Yagi at 50.40 MHz.

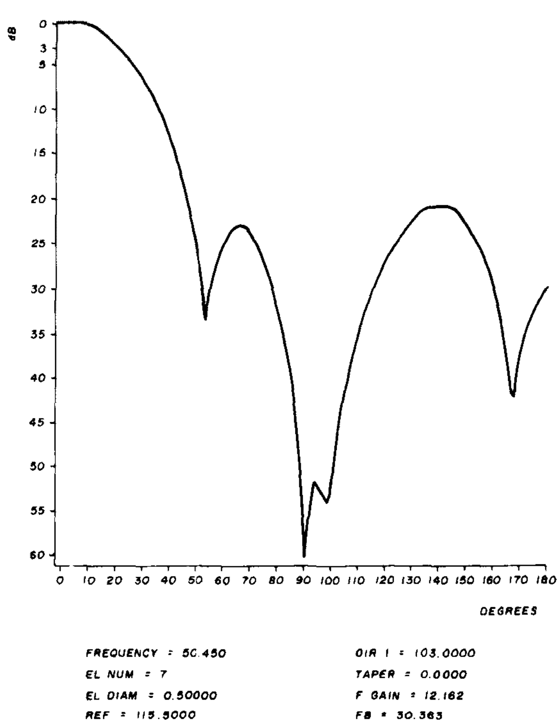

fig. 12. F/B optimized Yagi at 50.45 MHz.

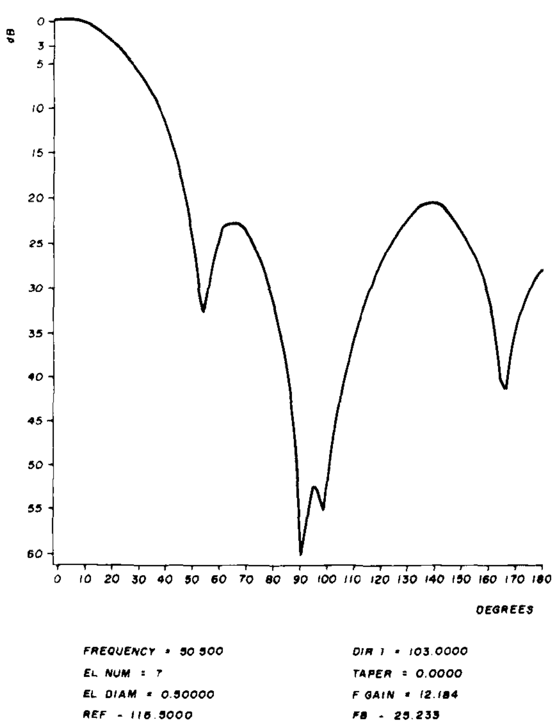

fig. 13. F/B optimized Yagi at 50.50 MHz.

table 7. Detailed performance parameters for the F/B optimized Yagi with a 0.000 taper.

frequency (MHz)	gain (dBi)	F/B (dB)
50.00	11.933	27.145
50.05	11.961	28.924
50.10	11.988	31.168
50.15	12.014	34.207
50.20	12.040	38.924
50.25	12.066	49.580
50.30	12.091	45.266
50.35	12.115	37.304
50.40	12.139	33.170
50.45	12.162	30.363
50.50	12.184	28.233

longer, these Yagis provide far more additional gain than this miniscule difference in boom length can explain. The model calculates 11.80 dBi gain for the NBS Yagi, as compared to 12.493 dBi for the gain optimized Tilton antenna, and 12.066 dBi for the F/B optimized Tilton antenna.[8] With approximately 19 dB of F/B, the NBS Yagi is superior to the gain optimized Tilton Yagi, but falls far short of the F/B optimized Tilton Yagi.[9] *It should be noted that* F/B *was never a design* criterion for any of the NBS Yagis. In conclusion it can be stated that as a result of the calculated performances of either of these two Tilton/Greenblum Yagis, either is preferable to the 1.2 wavelength NBS Yagi. The F/B optimized Yagi is the more preferable of the pair. The additional director on what is substantially the same boom length, provides a significant increase in performance.

Next month's installment addresses the general subject of Yagi performance optimization. In addition to discussing approaches taken by authors in the engineering literature, specific examples of techniques from the Amateur Radio literature will be modeled and iterated.

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ham radio

Attention HFer's — Don't skip Joe's column this month. The principles he discusses apply to other frequency ranges as well. — **Editor**

VHF/UHF WORLD

Joe Reisert W1JR

the VHF/UHF primer: an introduction to filters

Filters seem to be one of the least understood subjects in Amateur Radio despite the fact that numerous articles have been written on the subject. Have you ever heard an Amateur say, "I have this great 2-meter filter with only a 1-MHz bandwidth" (while failing to note or be aware of the fact that the insertion loss of the filter is 3 dB) or worse, say "I have this 70-cm (432 MHz) filter with only 0.15 dB loss," forgetting to mention that the bandwidth of the filter is 100 MHz!

Although most Amateurs seem to understand the difference between the categories of filters, described later in this column, they seldom understand how filters work. Sometimes they fail to recognize the interrelationship between bandwidth, insertion loss, and out-of-band attenuation, and what the best design for an application may be or how one goes about designing and building a filter. Although space doesn't permit presentation of a detailed design compendium, some designs and guidance will be provided in this and future columns.

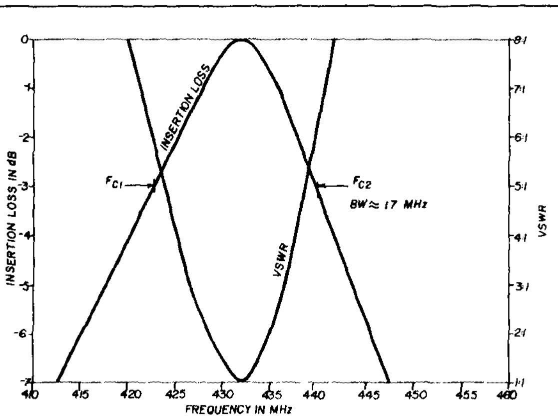

fig. 1. Insertion loss and VSWR of a 70-cm cavity filter.

filter basics

Filters can be classified according to two general types: absorptive and reflective, with absorptive filters the less common type. Absorptive filters accept all frequencies received. The desired frequency or frequencies are passed through to the output, while the undesired frequencies are directed to either a *built-in or external load* where they are dissipated. An example of an absorptive filter is the diplexer for terminating a double balanced mixer recommended in a previous column.[1] The better homebrew TVI lowpass *filters are absorptive* filters.[2] The advantage of this type of filter is that the source or transmitter always sees a good VSWR almost regardless of frequency, harmonics, etc.; its disadvantages lie in the fact that the filter is usually twice as complex and requires an extra termination.

The most common type of *filter is* the reflective type, which allows the desired signals to pass through from the input or source to the output, but reflects the undesired frequencies back to the source. A good analogy to this type of filter is the typical Amateur Radio antenna system. The resonant (or bandpass) frequency of the antenna has, if properly matched, a low

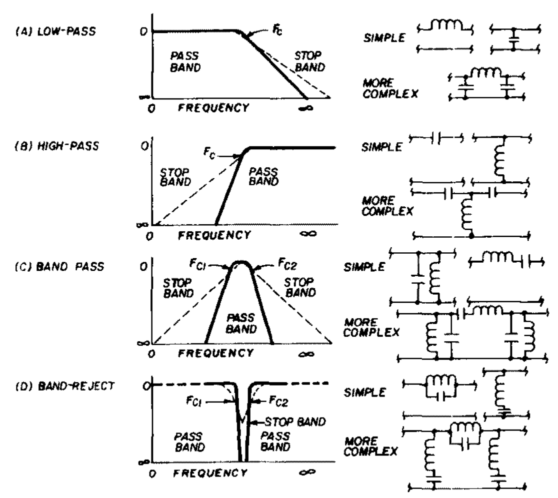

fig. 2. Basic categories of reflecting type filters in the frequency domain, illustrating typical simple and more complex circuits as well as frequency responses. Dotted responses represent simple filters, while solid lines indicate more complex filters.

VSWR, and radiates power into space. Frequencies off resonance are usually reflected back to the transmitter in the form of a high VSWR.

Fig. 1 illustrates this characteristic. A typical frequency versus input to output amplitude response and VSWR for a 70-cm reflective type bandpass filter is shown.[3] Note that the insertion loss, as expected, increases on either side of the center frequency, F_O. Note, too, that the VSWR rapidly increases on either side of F_O in a similar manner, and is approximately 6:1 at the half power or 3 dB down cutoff frequencies, F_{c1} and F_{c2}.

filter categories

Within each of the two filter types, absorptive and reflective, there are four basic categories: low-pass, high-pass, bandpass, and band reject or band stop. Each has a specific passband in which the insertion loss is low and a cutoff frequency (or frequencies) where the output is one half or 3 dB lower than the power in the desired passband. Sometimes filters consist of a combination of two or more of these categories. For instance, a low-pass and a high-pass filter may be connected in cascade to form a bandpass filter.

Low-pass filters have low loss up to the cutoff frequency and high insertion loss above this frequency (**fig. 2A**). They are most often used on the output of oscillators and transmitters to prevent harmonics from being radiated. (Typical low-pass filter schematics are also shown.) As the number of elements in the filter is increased, the passband insertion loss and the stop band attenuation increase. Also the passband insertion loss approaching cutoff is less as shown.

High-pass filters have high insertion loss below and low loss above the cutoff frequency (**fig. 2B**). They are most often used to prevent lower frequency transmitters from saturating the front end of a receiver. A common application involves using the filter on TV set inputs to prevent fundamental overload and on the input of an HF receiver to prevent overload from high power broadcast band transmitters. As the number of elements in the filter increases, the passband insertion loss and the stop band rejection increase, while the passband insertion loss approaching the cutoff frequency is less as shown in **fig. 2B**.

Bandpass filters have low insertion loss between two cutoff frequencies and high loss above and below the cutoff frequencies (**fig. 2C**). They are probably the most common form of filter used by VHF/UHF Radio Amateurs and can be considered as a combination of a low-pass and a high-pass filter. They are most often used as front-end filters to reject out-of-band signals that may overload a receiver. As the number of sections increases, the passband insertion loss and the stop band rejection increase, but the passband insertion loss is also less as you approach the cutoff frequency.

Band reject or band-stop filters have high insertion loss at a specific frequency or band of frequencies but low loss on all other frequencies (**fig. 2D**). They are used to eliminate discrete frequencies such as an IF image, a birdie, or local transmitter. For increased rejection, the number of sections must be increased.

filter parameters

The most important filter parameters are usually bandwidth and insertion loss. Secondary considerations are VSWR, passband ripple, out-of-band rejection, and shape factor. These parameters are all interrelated.

Insertion loss is especially important when a filter is at the input of a low-noise receiver because filter insertion loss at this location increases the noise figure by the same amount. If a filter follows a transmitter, the output power will be reduced by the amount of the insertion loss and if the filter is too lossy, it may break down or be destroyed when subjected to high power. Generally speaking, insertion loss increases if either the filter bandwidth is decreased, the number of sections in the filter is increased or the unloaded Q, Q_u, of the inductors is low (more on this later in this column).

Bandwidth is very important because it defines the operational frequency range over which signals will not be attenuated more than 3 dB. Bandwidth should never be any less than necessary, since narrow bandwidth usually goes hand-in-hand with higher passband insertion loss and more critical tuning.

VSWR is usually low at the center

frequency of a well designed and built reflective type bandpass filter and climbs abruptly near the cutoff frequency going toward infinity in the rejection band (**fig. 1**). VSWR will also be low in the passband of other categories of filters (to be discussed later in this article) but will increase abruptly as the cutoff frequency is approached. If a low VSWR is desired over a wide band, either the bandwidth of the filter must be increased or additional filter sections added.

Passband ripple is a function of the design parameters, the insertion loss and how well the filter is tuned and built. Remember that at each point where there is ripple, the input VSWR will increase or decrease accordingly. Ripple is associated with certain classes of filters as shall be discussed shortly.

Out-of-band rejection is a function of the design parameters, the bandwidth, the types of components used in the filter, construction, and alignment; but most importantly, it is primarily determined by the number of elements or tuned sections in the filter. If a low out-of-band rejection filter is used, it could mean overload to a low-noise receiver or harmonics radiation from a transmitter. Sometimes the bandwidth cannot be sufficiently reduced, and the number of sections in the filter must be increased to adequately attenuate out-of-band signals. As mentioned earlier, when the number of sections in the filter is increased, so is the insertion loss. Consequently, to increase out-of-band rejection, the filter complexity will usually increase.

Similarly, the shape factor of a filter is a direct function of the number of elements or tuned sections in a filter as well as the design parameters. VHF/UHF filters usually do not approach rectangular shape factors like IF filters because losses and the complexity of the filter would be difficult to work with at these frequencies. Hence, if good shape factor is required, the desired frequency range is usually converted to a lower frequency where insertion loss and components do not provide such a constraint.

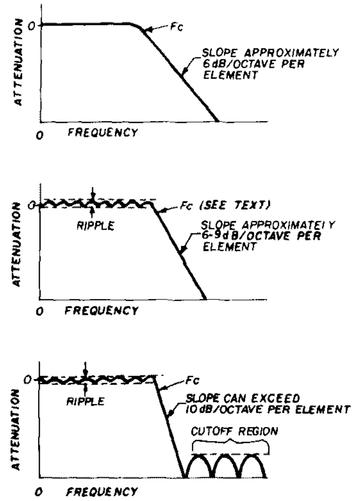

fig. 3. Typical amplitude and frequency responses of the different classes of low-pass filters.

filter classes

Frequency domain filters are the most commonly used "LC" filters. The three classes most often used are the Butterworth (or maximally flat), the Chebyshev, and the Cauer (or elliptical function). Each has specific characteristics defined by the design equations used to determine the component values and the tuning of the filter.

The most common filter, the Butterworth, or maximally flat filter, has been around for a long time. Its main advantages are that it has low insertion loss, low VSWR, a flat passband response and standard design tables are available.[4-7] Designs using normalized tables from these references are based on the 3 dB cutoff frequency. The disadvantage of the Butterworth filter is that its attenuation is only moderate out-of-band. A typical low-pass Butterworth filter response is shown in **fig. 3A**. Note that the attenuation in the stop band increases approximately 6 dB per octave (2 times the frequency) per filter element.

Chebyshev filters are a result of the development of modern filter theory. They are most often used when steeper stop band attenuation (than Butterworths can provide) is required. For example, the stop band attenuation for a properly designed Chebyshev filter may increase (depending on passband ripple selected) to approximately 9 dB per octave per element, up to 3 dB more than a Butterworth design. The principal drawback of a Chebyshev filter design is that it has ripples in the passband. The greater the desired attenuation out-of-band per number of elements in the filter, the more the ripple in band. As a point of interest, when the passband ripple goes to 0 dB, the design goes from Chebyshev to Butterworth. Design tables in references 4, 5, 6, and 7 are available. Beware, however; some authors specify the cutoff frequency as the 3 dB down bandwidth,[5] while others specify the cutoff frequency as the limit where the ripple bandwidth is constant.[6,7] A typical Chebyshev low-pass filter frequency response is shown in **fig. 3B**.

Cauer, or elliptical function filters, are also a result of the development of modern filter theory. Similar to the Chebyshev, they have resonant circuits that are used to produce notches in the stopband and are most often used when very steep attenuation is required just outside of the passband, particularly when you want to notch a specific frequency or frequencies. Other applications include designs that require only a finite amount of stop

band attenuation. The passband ripple is similar to Chebyshev designs. However, in the stop band, they also have ripple and reach only a minimum level of out-of-band rejection determined by the design parameters and filter alignment. Elliptical function filters also require extra tuning elements and possibly some peculiar component values. Hence, they are usually only used at audio through HF. A typical low-pass elliptical function filter response is shown in **fig. 3C**.

physical characteristics

Filters are made in many shapes and sizes. As discussed earlier, when the desired stop band rejection increases, the number of components and sections in the filter must increase. At lower frequencies (audio and HF), discrete components such as disc capacitors and inductors are usually used. However, as the frequency increases into the VHF/UHF range, other types of components such as air variables and rods for inductors are used. Also, at VHF/UHF, filters may take on other physical structures such as cavity, microstrip, stripline, combline, interdigital, or helical.

A **cavity filter**, especially the quarter-wave type (**fig. 4A**) is quite common at UHF, is basically an enclosure, usually a cylinder, with a rod typically close to a quarter-wave in length down the center of the enclosure that takes the place of a discrete inductor. By making the rod shorter or longer, it can be resonated, if desired, without a tuning capacitor. This is often referred to in the microwave community as the TEM (Transverse Electromagnetic) mode. However, the rod is often shortened slightly and tuned with a capacitor, typically two metal plates whose spacing can be varied, at the top of the rod to facilitate "tweaking" the filter to frequency. Cavities have low loss if they are large in diameter (between 0.05 and 0.33 λ), close to 77 ohms in impedance (3.6:1 ratio between the inside diameter of cavity and the outside diameter of the rod), have good conductivity particularly at the high current point at the base of the rod, and especially if they are silver plated internally.

However most people ignore or are not aware of the fact that a quarter-wave cavity filter is also resonant at several other frequencies in addition to the design frequency. If the filter is tuned mainly by adjusting the length of the rod, the other resonances will be close to 3, 5, 7 (etc.) times the design frequency. Hence they are not good harmonic filters! These other resonances can be pushed higher and placed above the harmonic by foreshortening the quarter-wave rod and using capacitance tuning as just described. The shorter the rod, the higher the frequency of the other resonances.

Another favorite Amateur filter configuration is the **"half-wave"** type (**fig. 4B**). Basically composed of two quarter-wave filters in tandem, its chief advantages are its requirement of only one tuning capacitor and its very symmetrical out-of-band rejection (more on this later in this article). The inductor can even take the form of a flat strip, thus yielding a mechanical advantage in some situations, since the input and output are at different ends of the filter.

Interdigital filters are very common especially at UHF and above.[7-9] They get their name from their physical appearance — they look like fingers joined (or interdigitated) together (**fig. 4C**). The rods are nearly a quarter-wave long, and ground is alternated from side to side. The spacing between resonators, the rod diameters, and the thickness of the structure are the major factors in determining bandwidth. Because the open ends have some fringing capacitance, it is best to shorten the rods slightly and add a small tweaker such as a silver-plated screw. This type of filter, properly constructed, usually has low passband insertion loss and is easily duplicated. It is readily scaled in frequency by keeping the thickness, rod diameter, and spacing constant, and just changing the length. The new frequency will have the same percentage bandwidth as the original frequency. This type of filter is usually large and also suffers from the overtone problems as mentioned above with the quarter-wave cavity type filter.

Combline filters — so named because they look like the teeth of a metal comb — are very common in the industry, (**fig. 4D**). Their rods or resonators are usually about one-eighth wave long (in contrast to the quarter-wave interdigital type). These filters are usually used where space is at a premium. Frequently the resonators are close together, and partitions or walls are placed between them, with the height of the partitions determining the coupling and hence overall filter bandwidth. The advantage of this type of filter configuration is that if the resonators are kept short, the response to overtones (as discussed above) will be much higher in frequency and may not be detrimental, as in the case of the interdigital type. One advantage or disadvantage, depending on your point of view, is that tuning capacitors are required.

Microstrip and **stripline** techniques are often used, especially when designing combline and interdigital filters. The spacings and impedances are set for the desired filter response. Losses in this type of construction are frequently higher than those using resonators etc. but this type of structure is usually reproducible.

Let us not forget **helical resonator filters**,[10,11] which resemble a large coil, usually with large diameter wire wound like a helix antenna (**fig. 4E**). Placed in a structure similar to the cavity filter, a helical resonator filter is usually used when relatively high inductance and low passband insertion loss are required, especially on the VHF and lower UHF frequencies.

filter anomalies

Several things must be understood before filter design can begin. It is assumed at the outset that all formulas used and the computed values are correct; the principal filter design problems occur with the values of the components and the types of components selected. Components are usually

chosen using standard design equations or normalized tables.[4-7,12] It would be very difficult to build a filter if the component values were not standard or readily available. Therefore, it is usually preferable to make some of the components variable, especially in bandpass filters. I prefer to use variable capacitors instead of variable inductors because they usually have higher Q_u and are easier to tune.

If insertion loss is important, and/or if the bandwidth of the filter is narrow, the components chosen must have high Q_u at the filter frequency. Air variables or small tuners made from plated screws are usually preferred at VHF/UHF frequencies. Inductance in series with the capacitor may cause loss or distortion such as a decrease in attenuation at some frequency or frequencies in the rejection band.

However, the greatest filter loss usually occurs in the inductors. The Q_u of an inductor should be as high as possible. For reference, I have shown some typical values of Q_u in **table 1**. This table is by no means complete, but can be used as a guide. Note that most discrete inductors have only moderate Q_u while cavities and helicals are high.

To determine the loss of an inductor in a bandpass filter, it is first necessary to determine the loaded Q, "Q_ℓ" of the filter as follows:

$$Q_\ell = \frac{F_o}{F_{c2} - F_{c1}} \qquad \textbf{(1)}$$

where F_o is the center frequency, F_{c1} is the lower cutoff (-3 dB) frequency and F_{c2} is the upper cutoff frequency all in the same units. For example, the filter in **fig. 1** has a center frequency of 432 MHz and upper and lower cutoff frequencies of approximately 440 and 423 MHz respectively. Therefore:

$$Q_\ell = \frac{432}{(440 - 423)} = 25.4$$

The loss in a filter is directly related to the ratio of the Q_u and Q_ℓ of the filter as shown in **eq. 2**:

$$\textit{insertion loss (dB)} = 10 \log_{10} (1 - Q_\ell / Q_u)^2 \qquad \textbf{(2)}$$

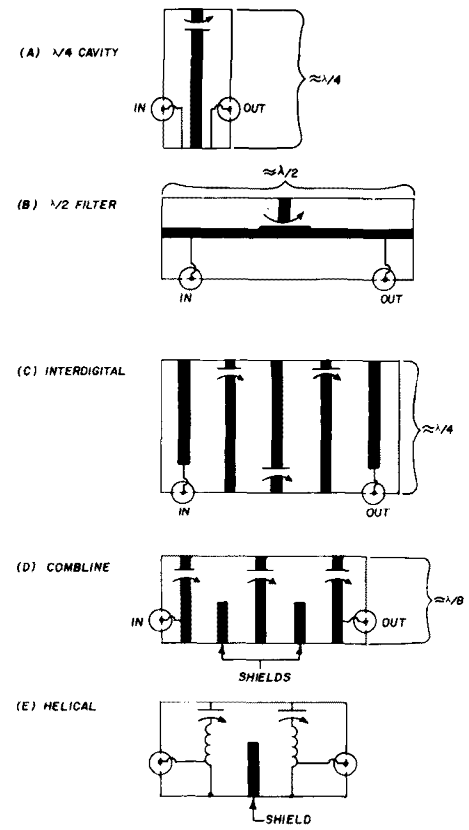

fig. 4. Typical filter geometries.

table 1. Typical Q_u versus frequency for some commonly used inductors.

inductor type	frequency range (MHz)	typical unloaded Q
toroid	1- 100	50- 200
0.25″ (6.35 mm) diameter coil	50- 500	300- 400
0.5″ (12.7 mm) diameter coil	50- 300	400- 500
microstrip (on G-10 PC board)	400-2300	175- 420
microstrip (on PTFE fiberglass)	400-2300	200- 775
1″ (25.4 mm) diameter 77-ohm cavity	300-1000	500-1000
3″ (76.2 mm) diameter 77-ohm cavity	100- 500	1000-1500
1″ (25.4 mm) helical resonator	100- 500	500-1000

For the sake of simplicity, I have drawn a graph for the most common ratios of Q_u/Q_ℓ in **fig. 5**. If the same example is used from **fig. 1**, and the Q_u of the inductor is approximately 1000 (from **table 1**), the Q_u/Q_ℓ ratio is 39.37, and the insertion loss of this filter will be approximately 0.22 dB. Now if the inductor had a Q_u of only 300, the Q_u/Q_ℓ ratio would be 11.8, and the loss would rise to approximately 0.77 dB — a huge increase! *This example shows why cavities are preferred at VHF/UHF frequencies.*

Before leaving this subject, it might be wise to explain how insertion loss can be determined in a multi-section filter. Cohn provides an equation to

determine this loss.[13] I have simplified his formula and listed the values or "K" factor necessary to determine the loss for common two through seven-section Butterworth bandpass type filters. Other types of filters may have slightly higher insertion losses. To obtain the loss of a multi-section bandpass filter, use the following equation:

$$\text{total insertion loss (dB)} = K(Q_\ell / Q_u) \quad \textbf{(3)}$$

where K is the number from **table 2** for the number of sections in the filter and Q_ℓ and Q_u are as determined above. For example, if the filter in **fig. 1** had 3 sections but the same Q_u (1000) and Q_ℓ (25.4), the insertion loss would be approximately 0.441 dB, much greater than the single section. If the Q_u were only 300, the insertion loss would climb to 1.47 dB. Hence, you are trading insertion loss for better *out-of-band rejection* — not always a bad compromise!

Insertion loss has one other detrimental effect. Since the insertion loss as you pass through the filter is cumulative, the amount of energy reaching each successive section in the filter is less. As a result, if the insertion loss of a filter is high, the bandwidth of the filter and its ripple characteristics may change from those predicted or calculated. Hence it is best to design a filter with slightly wider bandwidth than required because losses reduce bandwidth when the filter is finally tuned to frequency.

table 2. Approximate insertion loss in a multi-section Butterworth bandpass filter can be determined by using this information in conjunction with *eq. 3*.

number of sections	K factor
2	12.28
3	17.36
4	22.70
5	28.10
6	33.55
7	39.10

Other anomalies depend on the topography chosen. For instance, if the coupling into and out of a bandpass filter is capacitive, the filter will acquire a high-pass characteristic in the rejection band. Inductive coupling, a low-pass characteristic, may also occur. A combination of the two will yield a more symmetrical rejection band. This is illustrated in **fig. 6**.

Finally, input or output VSWR due to component selection, tuning, or loading will cause increased insertion loss, asymmetry in the pass or reject band and/or ripple in the passband. In this regard, bandpass filters with single sections are preferred, especially when they are placed ahead of a low-noise preamplifier since the point of minimum insertion will usually fall at the center of the band. Hence, if there is a severe mismatch (typical of low-noise amplifiers) the chances are that the minimum ripple will remain at the center frequency. To a somewhat lesser degree, 3 or 5-section filters are also preferred over 2 and 4-section filters for the same reason.

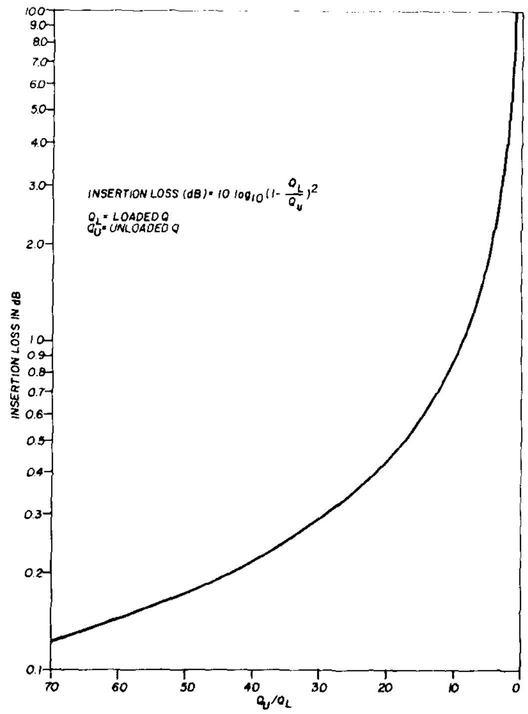

fig. 5. Insertion loss is a function of the loaded and unloaded Q of the filter.

designing filters

Low and high-pass filter designs are available in references 4 through 7 and also in 12. Band reject or band stop filters, primarily used in diplexers for FM repeaters where very high attenuation is needed between two close frequencies[14] are documented in references 4 and 5. My favorite bandpass filter design procedures are found in Zverev's book (reference 5). Some selected Chebyshev bandpass filter designs can be found in Anderson's *ham radio* article of June, 1977.[15] Design programs for many of these filters are now available for computer-oriented[16] Amateurs.

Bandpass filters seem to be the most widely used by VHF/UHFers, especially in receivers, transmitters, and ahead of preamplifiers. Hence I have dedicated the major part of this column to that subject and direct you to the applicable reference for the other categories of filters because they are usually easier to design. However many of the suggestions in this column apply to filters of any category.

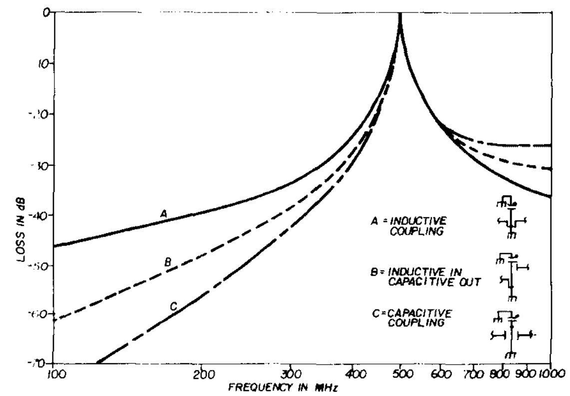

fig. 6. The frequency response of a 500-MHz cavity filter depends on coupling technique.

adjusting and measuring performance

This is really a subject for a whole column in itself. Remembering what has been said so far, the simple single-section filters (such as quarter-wave cavity) can usually be adjusted to frequency simply by placing a good low VSWR 50-ohm termination on the output of the filter and tuning for mini-num VSWR. Multi-section filters usually require some sort of sweep set-up with detectors. A typical procedure s to first align the filter for approx-mately the amplitude response ex-pected. Then the final testing and alignment is performed by observing VSWR over the entire bandwidth.

Dishal used a slightly different meth-od, in which he adjusted a slotted line and successively either shorted or opened each section of the filter as it was adjusted.[17] Suffice it to say that he proper alignment of a multi-section ilter requires both excellent test equip-nent and the skill to recognize what s taking place.

summary

It is important to recognize the dif-erent electrical and physical properties n order to chose or understand what ilter is the type required for a specific pplication before designing or build-ng it. It is also nice to at least know he difference between a Butterworth, Chebyshev, or Cauer filter, and whether it is a low-pass, high-pass, band stop or bandpass type. Hopefully this article has provided sufficient information to make this possible.

acknowledgements

I would like to thank Ron Matthews and Keith Whynot, WA1GZN, for their helpful suggestions in preparing this month's column.

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important VHF/UHF events in August, 1984

August 5-6: *ARRL UHF Contest*
August 11: *1945 UTC, predicted peak of Perseids meteor shower*
August 28: *Moon at perigee*

ham radio

short circuit

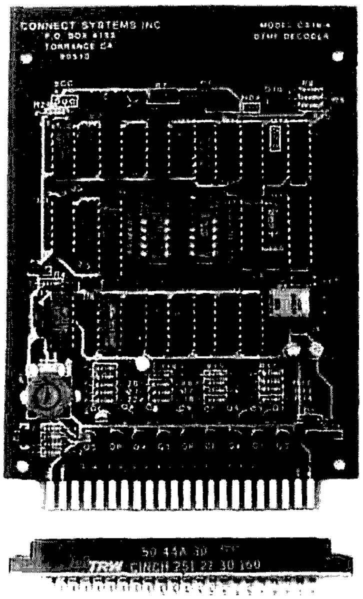
digital audio filter

In WIØER's "A Digital Audio Filter for CW and RTTY" (August, 1983, page 61), U4 and U5 should be labeled LF356N, not LF365N.

NEW products

touch-tone decoder

Connect Systems, Inc., has introduced a new 16-function touch-tone decoder board. Designated model CS-16, the decoder will securely control virtually any apparatus via radio or wireline. The CS-16 is especially useful for controlling various repeater on/off functions.



A unique feature of the CS-16 is dual password control. Two separately user-programmable three-digit passwords create heirarchy control capability. The primary control password can access all 16 of the available functions. The secondary password, however, can access only 8 of the 16 functions. A special primary password command, capable of enabling or disabling secondary password access, is available. The CS-16 provides such a high degree of multi-level security that control can be accomplished directly on voice channels, thus eliminating the need for separate control frequencies.

The CS-16 provides 16 independently controllable on/off latched functions. Each function is provided with an open collector and 5-volt CMOS logic output. A strobe output is also available in open collector and logic format. This output can be used to gate repeater audio so that DTMF control commands are not retransmitted.

A power-up reset feature causes all outputs to be in the "off" state after application of power. An audio pre-amp with level control permits the crystal controlled tone decoder to operate over the wide input range of 10 μV to 2 volts. A strobe LED lights when any of the 16 buttons on a pad is pressed. (The CS-16 can also be used with 12 button pads). An on-board voltage regulator permits operation with a 10-25 VDC power source. The CS-16 incorporates reverse polarity protection and draws less than 20 mA from the supply.

For more information, contact Connect Systems, Inc., 23731 Madison Street, Torrance, California 90505.

Circle #301 on Reader Service Card.

extended warranty on satellite TV equipment

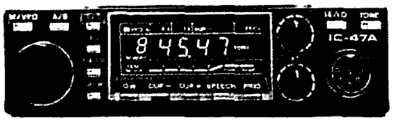
The R.L. Drake Company has announced that it has extended the limited warranty on the new Drake ESR-240 satellite earth station receiver and all other Drake satellite television equipment from one year on parts and 90 days on labor to one year for *both* parts and labor because no problems have been experiened with the new receiver's state-of-the-art infrared tuning feature.

For further information, contact R.L. Drake Company, 540 Richard Street, Miamisburg, Ohio 45342.

two new mobiles from ICOM

ICOM has added two more transceivers to its line of ultra-compact mobiles: the IC-27H 45-watt 2-meter mobile and the IC-47A 25-watt 440 MHz mobile.

Standard features of the IC-27H include 45 watts output, compact size (1-5/8" high × 5-1/2" wide × 9-3/8" deep), a built-in internal speaker for easy mounting, nine full-function memories, 32 built-in PL frequencies, professional communications design and styling, and an IC-HM23 DTMF microphone with up/down buttons. Scanning functions include memory scan, band scan, and priority scan. An internal lithium memory battery backup maintains memories for up to five years.



The features of the IC-47A are similar to those of the IC-27H, with the exception of 25-watt operation and somewhat smaller size.

Both units include the IC-MB27 mobile mount. A variety of options, including an IC-UT16 speech synthesizer and IC-SP4 and SP5 external speakers, are available for both units. The IC-27H is priced at $409; the IC-47A, at $469.

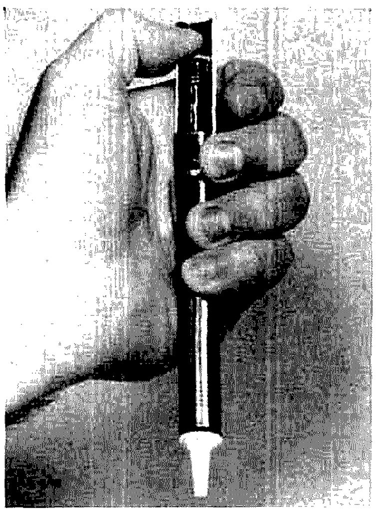
For further details, contact ICOM America, Inc., 2112 116th Avenue, N.E., Bellevue, Washington 98004.

Circle #302 on Reader Service Card.

desoldering pump

A compact spring-powered desoldering pump has been introduced by the Ungar Division o Eldon Industries, Inc.

The Ungar 7870 desoldering pump can be operated with one hand, leaving the other free to hold the soldering iron. A spring-loaded pistor



is set with the thumb and released by pushbut ton. The vacuum created by a piston stroke c less than two inches instantly removes molte solder. The thumb tap is recessed into the har dle to prevent eye injury doing close-toleranc desoldering, and a plated interior rod cleans th tip each time the pump is used.

Further information is available from Ungai P.O. Box 6005, Compton, California 90220.

Circle #304 on Reader Service Card.

new terminal

Robot's new 800C specialty mode terminal an improved version of their Model 800 "supe terminal," provides Amateur Radio operators a all-in-one package with display, storage, an automatic operation for the transmission an reception of RTTY and Morse code signals. A major feature of the 800 is its built-in demodu lator, which uses separate active discriminato filters for the demodulation of the RTTY signa

Key features of the new 800C terminal includ a 1023 character transmit buffer, ten 64-charac ter message memories with soft partitioning, a RS-232 serial and Centronics parallel printer inte face, color SSTV graphics capability with eigh graphics memories (when used with Robot' new color scan converters), and battery back up on all memories. These new features are als available in a retrofit kit for existing model 800s

For more information, contact Robot Re search, Inc., 7591 Convoy Court, San Diego California 92111.

Circle #303 on Reader Service Card.

product REVIEW

Mini Jini Record Keeper

Mini Jini Record Keeper from Jini Micro Systems is a powerful, yet easy-to-learn, plug-in data-base manager for both the VIC-20 and Commodore 64. (The term "data base" is synonymous with "file cabinet," updated to take advantage of the latest in computer technology.)

In the hamshack we store important information in a number of different and sometimes disorganized ways: in logbooks, cardboard boxes, and drawers. This can all add up to a disorganized and difficult-to-use system of data management.

Mini Jini was originally developed for the PET line of Commodore computers as the JinSam 8.2 Data Manager. This software has been used extensively throughout the business world because of its powerful record keeping capabilities and its easy-to-use design.

NASA, for example, has made extensive use of JinSam in its management of landing site facilities for the space shuttle program. Space shuttles have two main landing sites: one at Cape Canaveral in Florida and the other in California at Edwards Air Force Base. Because of the possibility of problems that might preclude landing at either of these sites, NASA has designated five additional sites around the world. In order to be ready to use any one of these, and to hold costs down, NASA's contingency plan specifies that in the event of an unscheduled landing at any alternate site, equipment would be shipped to the landing site from a central location. Obviously, manual file maintenance of such an elaborate system would be costly and cumbersome. Using their Commodore CBM and the JinSam 8.2, NASA's records now contain fields (i.e., files) for equipment nomenclature, serial number, and present location. In just a few minutes, NASA personnel can do full file searches to produce organized shipping lists of equipment for each or all landing sites.

Since the developers of the JinSam 8.2, Jim and Nancy Iscaro, are both hams, it was only a matter of time before they turned their attention to using JinSam for Amateur Radio purposes. Noting that both the Commodore 64 and VIC-20 are quite popular in the Amateur Radio field, the Iscaros set about the task of converting JinSam to a usable format for the C-64 and VIC-20. The result was Mini Jini. Of tremendous interest is that this program is not limited to Amateur Radio use, but can be used to store a variety of business, household, or personal information. For the Radio Amateur, Mini Jini can be used to log QSOs, print QSL labels, inventory equipment, keep contest logs, organize magazine files, and catalog foreign phrases, to name just a few of its many uses.

Mini Jini comes in a manufactured board that is inserted into either the VIC-20 or C-64 cartridge slot. With a stock (unexpanded) VIC-20, you can store up to 50 full records of 250 characters each. With a 24K memory expander added, the VIC-20 can handle up to 500 records in memory. The C-64 will hold up to 750 records. For permanent file storage, it's necessary to add either a disk drive or cassette recorder.

The well written and informative instruction book makes data entry easy. Mini Jini's manual should answer just about any question you may have; it's also full of helpful hints and tips on how to get the most of Mini Jini's capabilities.

In the ham shack, computers are no longer a luxury. In fact, after seeing how many VIC-20 and Commodore owners responded to our reader survey (see September, 1983, for the survey, and January, 1984, for our editorial response), it's hard to imagine that there's *anyone* who doesn't have a computer to use! Record keeping with computers can be a real plus when you're contesting, QSLing, or organizing your collection of magazine articles. Mini Jini will do it for you with a minimum of fuss and hassle.

For more information, contact Jini Micro-Systems, Box 274 Kingsbridge Street, Riverdale, New York 10463.

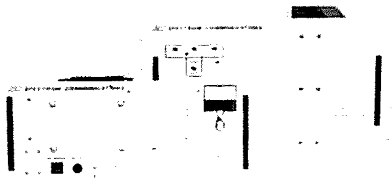
Circle #305 on Reader Service Card.

N1ACH

NEW products

power and power supply

The Spectrum SCA100V is a new 150-watt repeater/base station amplifier that operates in the 136-174 MHz range. Its unique heatsink and high efficiency cooling system design allow cool operation even under continuous duty conditions in a hot environment. The "behind the panel" heatsink permits use in a locking front door cabinet without loss of cooling effectiveness. It also features automatic high-VSWR shutdown/"bypass" with 4X auto-reset circuit and automatic amp bypass (if the power supply should fail or if the amp should overheat), as well as unusually tight RF shielding and heavy-duty construction. A 100-watt UHF version is also available.



NEW products

The companion power supply for tl SCA100V amplifier, the SCP30, may also l used for any type of high-power amplifier, se vice bench, or industrial application requiring very heavy duty supply. Its output is 13.6 VD at 25 amperes, continuous, 30 amperes inte mittent.

For further details, contact Spectrum Cor munications Corp., 1055 West Germantov Parkway, Norristown, Pennsylvania 19401-961

Circle #306 on Reader Service Card.

break-in adapter

Design Electronics Ohio has introduced its f break-in amplifier adapter, the QSK-1500. D signed to mate currently available full break radios — such as *all* Ten-Tec units, the Kenwo TS930, Drake TR-5, the Yaesu FT1, FT980, a FT757, and the ICOM751 — to either comm cial or homebrew linear amplifiers. Installati of the QSK-1500 requires no internal modific tions to most transceivers or amplifiers; a min modification is necessary with Ten-Tec units SSB.

While the QSK-1500 was designed primar with the CW operator in mind, it will also fun tion on SSB and RTTY. High power Amtor now also possible with this unit.

The QSK-1500 uses ultra high speed PIN dio switching and has no clicking or annoying rela The unit is designed to handle 1500 watts RF in a 50-ohm load at 40 WPM. Insertion loss is le than 0.6 dB on receive. Maximum receiver li voltage is 3800 mV before the protective circ is activated. An in-line fuse lamp will trip out 7.5 watts RF. Control lines for keyer must be f positive, cathode keying lines only. Amplifi switch time is less than 800 microseconds. Tl units measure 3 × 6.25 × 3.75 inches (7.6 15.9 × 9.5 cm) control, and 3.2 × 6.6 × 9. inches (8.1 × 16.8 × 23.5 cm) RF unit, ar weigh 5 and 4 pounds (11 kg and 8.8 kg) respe tively. The unit is available for $279 from eith Universal Amateur Radio, Inc., 1280 Aida Driv Reynoldsburg, Ohio 43068, or DEO, 4925 S Hamilton Drive., Grovesport, Ohio 43125.

Circle #307 on Reader Service Card.

new 40-meter antenna

Telex/Hy-Gain has introduced the "Disco erer," a new series of 40-meter antennas deve oped for high-performance operations in r sponse to the effects of declining sunspot a tivity on the 10-20 meter bands.

The new series consists of several configur tions. The Discoverer 7-1 is a 45-foot (13.7 mete rotatable dipole that can be added to many e isting beam antenna installations and tuned either 30 or 40 meters.

The Discoverer 7-2, a two-element beam wi

a wind load of only 6 square feet (0.56 meter^2), ·equires only a 25-foot (7.6 meter) turning radius. n addition to high forward gain and front-to-back ratio, it maintains a broad bandwidth in ex-cess of 190 kHz with SWR below 2:1.

The Discoverer 7-2 can be further enhanced with the addition of a Director Kit, thereby creating a three-element beam. This almost doubles the front-to-back ratio and forward gain, which in turn almost doubles the E.R.P. of the two-element version. All this fits on a 35-foot (10.7 meter) boom. The relatively compact array does not require a heavy-duty tower, but can be safely installed on a less expensive medium-duty tower such as the Hy-Gain HG52SS.

The suggested list price for the Discoverer 7-1 is $195.00. The Discoverer 7-2 and the Director Kit are listed at $435.00 and $272.00 respectively.

For more information, contact Telex Communications, 9600 Aldrich Avenue South, Minneapolis, Minnesota 55420.

Battery Manager™

Designed with the communications specialist in mind, the Battery Manager™ from URDC Measurements, Inc., analyzes and conditions all common types of 2-way radio NICAD batteries for optimum field performance and extended life.

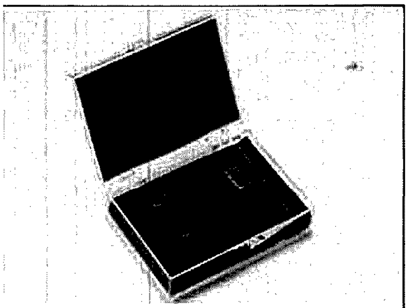
The unit is specifically designed to combat "Memory Effect" — the premature loss of power from a battery that's just been fully charged. Caused by repetitive shallow discharging followed by repeated overcharging, Memory Effect reduces reliability in the field and plays havoc with battery replacement schedules.

With the Battery Manager™, batteries that have been discarded as "not usable" can often be reconditioned and returned to service, and the life of batteries currently in use can be extended.

Battery Manager™ operates on 110V, 60 Hz and can be adapted at the factory for use on 20V, 0 Hz. Its maximum power consumption is 150 watts.

For more information, contact URDC Measurements, Inc., P.O. Box 880, West Jordan, Utah 84084.

Circle #308 on Reader Service Card.



New CMOS DTMF Chip Kit

Teltone's TRK-957 Kit makes it easier and less expensive to breadboard a low-power, central office quality DTMF detection system. All you need is a power source from 5 to 12 VDC. The sensitivity, wide dynamic range, noise immunity, and low-power consumption make the TRK-957 ideal for telephone switching, computer, and remote control applications. The TRK-957 DTMF Kit is only $24.75. To order call:

(800) 227-3800, ext. 1130.

TELTONE® ✓ 189

flea market

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COPY No special layout or arrangements available. Material should be typewritten or clearly printed (**not** all capitals) and must include full name and address. We reserve the right to reject unsuitable copy. **Ham Radio** cannot check each advertiser and thus cannot be held responsible for claims made. Liability for correctness of material limited to corrected ad in next available issue.

DEADLINE 15th of second preceding month.

SEND MATERIAL TO: Flea Market, Ham Radio, Greenville, N. H. 03048.

QSLs & RUBBER STAMPS — Top Quality! Card Samples and Stamp info — 50¢ — Ebbert Graphics 5R, Box 70, Westerville, Ohio 43061.

"KT5S" Super DX sloper 80-10M only $59.95. "KT5B" Multiband dipole 160-10M only $59.95. 2-KW roller inductor (28uh) $47.50. Weather boot kit (PL-259) $8.95 pp. Much more! Info available. Kilo-Tec, PO Box 1001, Oakview, CA 93022. Tel: 805-646-9645.

WANTED: Old RCA, Western Electric, tubes, speakers, amps. (713) 728-4343, Corb, 11122 Atwell, Houston, Texas 77096.

CUSTOM EMBROIDERED EMBLEMS — Enameled pins, your design, low minimum, excellent quality, low prices, free booklet. A.T. Patch Co., Dept. 65, Littleton, NH 03561. (603) 444-3423.

TRAVEL-PAK QSL KIT — Converts post cards, photos to QSLs. Stamp brings circular. Samco, Box 203-c, Wynantskill, New York 12198.

NH ALL-BAND QTH. South slope/retreat, Crotched Mountain, Greenfield. 4 room lodge, separate bunkhouse, 920' elevation, 1200' from highway and power mains, paved access, privacy, 59 acres. $95,000. Clarence Farr (W1WMK) Broker, Greenfield, NH 03047. (603) 547-2053.

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Greenside, Johannesburg
Republic of South Africa

RADIO ITEMS before 1930 wanted. Buying battery operated radios, horn and cone speakers, radio tubes and parts, radio literature — books, catalogs, magazines, radio advertising signs, posters. Gary Schneider, 6848 Commonwealth Blvd., Parma Heights, Ohio 44130.

QSL CARDS: $15.50/500, ppd. Free catalog. Bowman Printing, 743 Harvard, St. Louis, MO 63130.

DIGITAL AUTOMATIC DISPLAYS for FT-101's TS-520's, and most others. Six 1/2" digits. Write for information. Grand Systems, P.O. Box 2171, Blaine, Washington 98230. (604) 530-4551.

SELL: 40 year collection receiving tubes — dollar each. Antique molded mica transmitting capacitors, all sizes, voltages. Wanted to buy, machinist tools from retired toolmaker. W5QJT, PO Box 13151, El Paso, TX 79913.

RUBBER STAMPS: 3 lines $4.50 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

FREE catalog on new & used computers, hardware, keyboards, tapes, multiplexers, CRTs, etc. IN-X-Sales, Dept. B, Box 45, Tilton, NH 03276. (603) 286-3082.

"THE SWAP LIST" has bargains galore. Subscribe now! 6 months for $4.00; 1 year only $6.50. The Swap List, Box 988-H, Evergreen, CO 80439.

ROM-116 for TRS-80 RTTY with 5 program tapes and instruction manuals. 1.3 version. $200 or best offer. Ken Eaton, 6418 W. 78th St., Prairie Village, Kansas 66204.

HAM RADIO MAGAZINE Collection: Bound volumes 1969, 1970, 1971, 1972, 1973, 1974, 1975. In HR binders 1976, 1977, 1978, 1979. Loose copies 1980, 1981, 1982, 1983. Best offer for lot. W4UCH, P.O. Box 1065, Chautauqua, NY 14722. (716) 753-2654.

FOR SALE: Swan 350 transceiver w/ps $250. SA2040 antenna tuner $125. QF-1A audio filter $50. Kantronics CW/RTTY interface for Vic-20 with programs board + cables $135. Shipping included. Send money order. Package price $500. Write Jim Howell, KA4EBW, 18 Dan St., Salisbury, NC 28144. (704) 637-0313 evenings.

REPAIR, ALIGNMENT, calibration. Collins written estimates $25; non Collins $50. K1MAN (207) 495-2215.

MICROWAVE MIXERS. Two new Watkins-Johnson M77C X-band DBM's. $125 each. SMA or microstrip. Low noise, high isolation. Cost $382 each. John Perlick, 7435 Brunswick, Brooklyn Park, MN 55443.

CHASSIS and cabinet kits. SASE K3IWK.

NATIONAL RADIO CO equipment manuals price list SASE. Dust covers, NCX 3 or NCX 5 plus NCX A, pair $8.95 PP. Maximilian Fuchs, 11 Plymouth Lane, Swampscott, MA 01907.

ELECTRON TUBES: Receiving, transmitting, microwave . . . all types available. Large stock. Next day delivery most cases. Daily Electronics, 14126 Willow Lane, Westminster, CA 92683. (714) 894-1368.

HAM TRADER WEST "Pink Sheets" 25 cents a word for subscribers, $9 a year bimonthly, Canadians send U.S. funds, posted same day as U.S. from Canada. Others 50 cents a word, commercial 75 cents a word. Name address and phone free. Ham Trader West, Box 202, Lynden, Washington 98264.

WANTED: Cash paid for used speed radar equipment. Write or call: Brian R. Esterman, PO Box 8141, Northfield, Illinois 60093. (312) 251-8901.

$$$$ SUPER SAVINGS on electronics parts, components, supplies, and computer accessories. Free 40-page catalog for SASE. Get on our mailing list. BCD Electro, PO Box 830119, Richardson, TX 75083. Or call (214) 690-1102.

WANTED: *Old microphones, remote mixers other mic related* items. All pre-1935. Box Paquette, 107 E. National Avenue, Milwaukee, WI 53204.

DELTA LOOP BOOM for 10-20 meters. Lightweight, strong, quality boom with element brackets attached. Element kits available, or make up your own. Full details. D.L. Manufacturing, 485 Fifth Avenue, #1042, New York, NY 10017.

STERLING SILVER callsign jewelry pin or tack; $11.95, alligator tie clip; $19.95, all ppd. Info-SASE. Tom's Silver, PO Box 3758, Manchester, NH 03105.

RECONDITIONED TEST EQUIPMENT $1.00 for catalog. Walter, 2697 Nickel, San Pablo, CA 94806.

FOX-TANGO Newsletters -- Since 1972, the prime source of modifications, improvements, and repair of Yaesu gear, free to Club members. Calendar year dues still only $8 U.S., $9 Canada, $12 elsewhere. Includes five year cumulative index by model numbers, or send $1 for index and sample Newsletter. Fox Tango Club, Box 15944, W. Palm Beach, FL 33416.

RTTY-EXCLUSIVELY for the Amateur Teleprinter. One year $7.00. Beginners RTTY Handbook $8.00 includes journal index. P.O. Box RY, Cardiff, CA 92007.

IMRA International Mission Radio Assn. helps missioners — equipment loaned; weekday net, 14.280 MHz, 2-3 PM Eastern. Br. Frey, 1 Pryer Manor Rd., Larchmont, NY 10538.

"HAMS FOR CHRIST." Reach other Hams with a gospel tract sure to please. Clyde Stanfield, WA6HEG, 1570 N. Albright, Upland, CA 91786.

TENNATEST — Antenna noise bridge — out-performs others, accurate, costs less, satisfaction guaranteed, $41.00. Send stamp for details, W8URR, 1025 Wildwood Road, Quincy, MI 49082.

WANTED: Early Hallicrafter "Skyriders" and "Super Skyriders" with silver panels, also "Skyrider Commercial", early transmitters such as HT-1, HT-2, HT-8, and other Hallicrafter gear, parts, accessories, manuals. Chuck Dachis, WD5EOG, The Hallicrafter Collector, 4500 Russell Drive, Austin, Texas 78745.

VERY in-ter-est-ing! Next 4 issues $2. Ham Trader "Yellow Sheets", POB356, Wheaton, IL 60189.

FOR SALE: B&K model E-200D RF signal generator $280.00; electronic navigation model 310L 250kH-110 MHz RF linear amplifier $500.00; Leader model LSG-231 FM stereo signal generator $300.00; Eico model 150 solid state signal tracer $100.00; Lafayette Pip-Speak stereo speakers $50.00; Leader VLM model LP-70B 20k ohm per volt $40.00; Minolta SRT-200 35mm camera with UV filter, 50mm lens and case $125.00; RF Power Labs model V350 2 meter 350W linear amp. with fan cool $800.00; Monroe simplex autopatch, commercial quality, with various remote functions with Yaesu CPU-2500 2 meter memorizer radio $1250.00. TR-110 B&K Isopack isolator transformer $100.00. Mitchell Rakoff, 114-41 Queens Blvd., Suite 148, Forest Hills, NY 11375 (212) 591-0002.

Coming Events

ACTIVITIES

"Places to go..."

ALABAMA: The Huntsville Hamfest, Saturday and Sunday, August 18 and 19, Von Braun Civic Center, Huntsville. No admission charge. Exhibits, forums, air-conditioned indoor flea market and non-ham activities. Tours of the Alabama Space & Rocket Center available. A limited number of camp sites with hookups at VBCC. Reserved flea market tables available $4/day. Talk in on 34/94. For information: Huntsville Hamfest, 2804 S. Memorial Parkway, Huntsville, AL 35801.

WYOMING: The fifth annual High Plains Ham Roundup, sponsored by the Northern Colorado ARC, University of Wyoming ARC and Shy-Wy ARC, September 7, 8 and 9, Medicine Bow National Forest, Yellow Pine Campground, 35 miles west of Cheyenne. Campfire cookout and bring-your-own covered dish supper Saturday with singalong music and entertainment by regional talent. Barbequed hamburgers and refreshments provided by the committee. Giant tailgate swapfest, transmitter hunt and technical displays. No registration fee. Modest Forest Service charge for campers. Talk in on 22/82 and 25/85. For information: W7CGK, 1321 E. 22 Street, Cheyenne, WY 82001.

RADIO EXPO '84 sponsored by the Chicago FM Club, Saturday and Sunday, September 22 and 23, Lake County Fairgrounds, Rt. 120 & 45, Grayslake, IL. Major manufacturers and gigantic outdoor flea market. Flea market opens 6 AM. Exhibits 9 AM. Free parking and overnight camping. Reserved indoor flea market $5/day. Tickets $3.00 advance, $4.00 at gate, good for both days. Seminars, technical talks and ladies programs. Talk in on 146.16/76. SASE to Radio Expo '84, Box 1532, Evanston, IL 60204 or (312) 582-6923.

FLORIDA: The Platinum Coast Amateur Radio Society's annual Hamfest, September 8 and 9, Melbourne Auditorium Melbourne. Swap tables, meetings, forums, awards, tailgating For information or reservations: PCARS, PO Box 1004, Melbourne, FL 32901.

ONTARIO, CANADA: The Radio Society of Ontario's 16th annual Convention, October 5, 6 and 7, Westin Hotel, Ottawa Friday night eyeball and dance. Saturday and Sunday technical sessions, demonstrations and commercial exhibits. Saturday night banquet and dance. For information: RSO Convention Committee, PO Box 15806 Station "F", Ottawa Ontario K2C 3S7.

PENNSYLVANIA: The 47th annual South Hills Brasspounder and Modulators Hamfest, August 5, 9 AM to 4 PM, South Campus of the Community College of Allegheny County, Pittsburgh. Tickets $3 each or 2/$5. Indoor/outdoor flea market space available. Food and refreshments available. Free parking. Talk in on 146.13/73 and 146.52 simplex. For information: Jack B. Wood, 448 Jenne Dr., Pittsburgh, PA 15236

PENNSYLVANIA: Change of location for the Mid-Atlantic ARC Hamfest, August 12. (See announcement in July HR.) It will be held at the Bucks County Drive-in, Route 611, Warrington

MINNESOTA: The St. Cloud Amateur Radio Club's annual Hamfest, Sunday, August 12, 8 AM to 4 PM, Sauk Rapids M

ıcipal Park, Sauk Rapids Talk in on 146.34/94. For informa- on: St. Cloud ARC, PO Box 141, St. Cloud, MN 56302.

EW JERSEY: The 25th GCARC Ham/Compfest, sponsored y the Gloucester County ARC, August 26, 8 AM to 4 PM, loucester County College, Sewell. Admission $2.00 advance, 2.50 at door. Tailgating $3.00 per space. Seminars, contests, omputer demos, flea market, refreshments. Official VEC test- g center, testing Novice thru Extra. No pre-registration nec- essary; 610 forms available. Talk in on 147.78/18, 146.52 sim- ex, 223.36/224.96. For information or reservations: Mill Gold- an, K3WIL, 801 Crown Point Rd., Westville, NJ 08093. (609) 36-0500 or John M. Fisher, K2JF, PO Box 370, Pitman, NJ 3071 (609) 589-2318.

WA: The Des Moines Radio Amateur Association's Ham Computerfest, August 19, 9 AM to 5 PM, Veteran's Memorial uditorium, Des Moines. Donation $3.00 advance, $4.00 at oor. Expanded flea market, Amateur and computer dealers, onsignment tables. Refreshments available. Talk in on 16.22/82 and 440.5. For information and reservations: Bob ucker, KD0EO, PO Box 3711, Urbandale, IA 50322 (515) 76-4415 or Louis Seibert, N0ELI, 7515 Roseland, Urbandale, 50322 (515) 276-0272.

IAINE: The 1984 Windsor Hamfest, Saturday, September 8, Windsor Fairgrounds. Flea market, programs, speakers, dis- ributors, light meals and the traditional Saturday bean and asserole supper. Gate donation only one dollar. Camping vailable Friday and Saturday nights. Talk in on 146.22/82 epeater. For information: Don Hanson, N1AZH, RFD #2, Box 678, Greene, Maine 04236. (207) 946-7557.

IDIANA: The Tippecanoe Amateur Radio Association's 13th nnual Hamfest, Sunday, August 19, Tippecanoe County Fair- rounds, Teal Road and 18th Street, Lafayette. Grounds open t 7 AM. Tickets $3.00. Large flea market, dealers, refresh- ents and fun. Talk in on 13/73 or 52. For tickets or informa- on: Lafayette Hamfest, Route 1, Box 63, West Point, IN 7992.

ENTUCKY: The Central Kentucky ARRL Hamfest, spon- ored by the Bluegrass Amateur Radio Society, Sunday, ugust 12, 8 AM to 5 PM, Scott County High School, Longlick oad and US 25, Georgetown. Tech forums, awards, exhibits, C facilities. Free outside flea market space. Tickets $3.50 ~~dvance and $4.00~~ at gate. For information or tickets: Edward . Bono, WA4ONE, P.O. Box 4411, Lexington, KY 40504.

KLAHOMA-KANSAS State Line Area: Great Salt Plains 2nd nnual Hamfest, August 26, 9 AM to 5 PM, Community Build- g, south side of Great Salt Plains Lake. Tech forums, meet- igs, free swap tables, refreshments, Novice exams and noon otluck dinner. Overnight camping and RV hookups at Lakes tate Park. Talk in on 147.90-.30 Salt Plains Repeater. For formation: Steven Walz, WA5UTO, Box 222, Cherokee, OK 3728 (405) 596-3487.

EW JERSEY: The Ramapo Mountain ARC, WA2SNA, pre- ents its 8th annual flea market August 18, Oakland Amer- an Legion Hall, 65 Oak St., Oakland (20 miles from GW ridge). Admission $1.00. Non-ham family members free. In- oor tables $6.50. Tailgating $3.00. Talk in on 147.49/146.49 nd 52. For information: Tom Risseeuw, N2AAZ, 63 Page Dr., akland, NJ 07436. 337-8389 after 6 PM.

IISSOURI: The St. Charles ARC's Hamfest '84, August 26, t. Charles City Hall Complex. Barbeque provided by the Har- ester Lions. Nearby Riverfront Park and historic south main treet area. General admission $1.00. Talk in on 146.07/67 nd 52 simplex. For information: Ron Ochu, KO0Z, 1914 N. th St., St. Charles, MO 63301.

IRGINIA/WEST VIRGINIA: The Bluefield Hamfest, Computer Satellite TV Fair, Sunday, August 26, 9 AM to 3 PM, Brush- ork Armory-Civic Center, 1 mile north of Bluefield, West Vir- inia on US 52. Sponsored by the East River ARC. Admis- ion $3.00. Children under 12 free. Large indoor flea market nd other activities. Paved parking, food on site. Talk in on 44.89/145.49 and 146.52. For information: Don Williams, /A4K, 412 Ridgeway Drive, Bluefield, VA 24605.

ENNESSEE: The Short Mountain Repeater Club is sponsor- ig the Lebanon Hamfest, Sunday, August 26, Cedars of Leb- non State Park, US 231, Lebanon. Outdoors only. Bring your wn tables. Food and drink available. Talk in on 46.31/146.91. For information: Morris Duke, W4WXO, 210 ısspayne Drive, Donelson, TN 37214.

'ENNSYLVANIA: The Uniontown ARC (W3PIE) will hold its 5th annual Gabfest, Saturday, September 8, Club grounds, Ild Pittsburgh Road off Rt. 5 & 119 bypass, Uniontown. Free offee, free parking, free Swap & Shop with registration of 3.00 each, 2/$5.00. Good food at refreshment stand. Talk ı on 147.645-.045 & 144.57-.17. For information: John Cer- ıak, WB3DOD, UARC Gabfest Committee, PO Box 433, lepublic, PA 15475. (412) 246-2870.

'ENNSYLVANIA: The Central Pennsylvania Repeater ssociation's 11th annual Hamfest/Computerfest, August 26, lershey. Adjacent to "Hersheypark" Chocolate town USA. legistration $3.00. Wives and children free. Special reduced dmission to Hersheypark for families of registrants. Large ıdoor dealer and flea market area. 10' indoor spaces $8 each. ' tables $4 each. Single electric plugs $1 each. Large out- door tailgating area. Refreshments available. Talk in on 145.47, 146.76 and 146.52. For information: Timothy R. Fanus, WB3DNA, 6140 Chambers Hill Road, Harrisburg, PA 17111. (717) 564-0897 (Noon to 8 PM).

MICHIGAN: The Grand Rapids Amateur Radio Association's annual Swap and Shop, Saturday, September 15, Hudson- ville Fairgrounds. Indoor sales area and outdoor trunk swap area. Gates open 8 AM. Talk in on 146.16/76. For informa- tion: Grand Rapids ARA, PO Box 1248, Grand Rapids, MI 49501.

PENNSYLVANIA: The Tioga County Amateur Radio Club's 8th annual Hamfest, Saturday, August 25, Island Park, Bloss- burg. 9 AM to 5 PM. Flea market, dealers, traders, computer demo, QSL contest, transmitter hunt, XYL and Harmonic pro- grams, RC airplanes and more. Talk in on 146.19/79, 146.52/52 and CB. Admission $3.00. XYLs and kids free. For information: Carl E. Kimble, WB3EUE, PO Box 37, Cowan- esque, PA 16918. (814) 367-5345.

IOWA: The Iowa 75 Meter Net will sponsor a Hamfest and picnic, Sunday, August 26, W.K.W. Park, North of Hampton. A potluck meal at noon followed by a short program. Talk in on Mason City 146.15-.65. For information: WD0FWB or WB0JFF.

INDIANA: The Marshall County Amateur Radio Club's annual Hamfest, Sunday, August 26, 8 AM to 2 PM, Marshall Coun- ty 4H Fairgrounds, Argos. All kinds of radio gear. Dealer tables available — 8'/$5. Good food and drink. Ladies' activities. Talk in on 146.52 simplex. Tickets $2.00 advance; $3.00 at door. For information: Marshall County ARC, PO Box 151, Plymouth, IN 46563 or call Bob Nellans, KB9DE (219) 892-5224.

OPERATING EVENTS

"Things to do..."

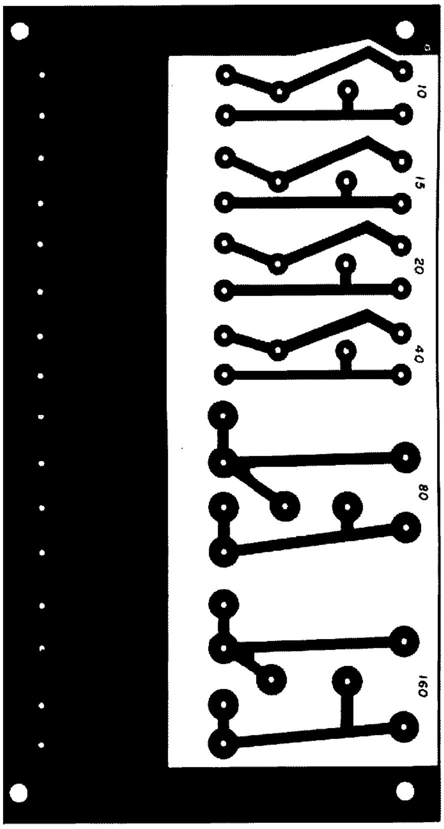
AUGUST 18-19: The Bergen Amateur Radio Association, Paramus, New Jersey, will operate special event station, K2TM, from 1500 to 2400Z to celebrate the Club's 21st anni- versary. Certificate for large SASE and QSL via K2UFM, 31 Forest Drive, Hillsdale, NJ 07642.

AUGUST 18-19: The Cascades Amateur Radio Society (CARS) in conjunction with the Michigan Space Center, Jack- son, Michigan, will operate WB8CSQ during Space Day ac- tivities. 0000 GMT August 18 through 1700 GMT August 19. For a special Space Day certificate, send log info and $1.00, to cover postage and materials, to CARS, Space Day '84, PO Box 512, Jackson, MI 49204.

SEPTEMBER 8: The West Alabama Amateur Radio Society (WAARS) will operate the 2nd annual special event station, KE4TN, from the campus of the University of Alabama, 1300Z to 2400Z, to commemorate the great college football coach, Paul 'Bear' Bryant. The Club will offer a handsome commem- orative certificate to any station worked. Send $1 and a large SASE to: West Alabama ARS, PO Box 1741, Tuscaloosa, AL 35403.

AUGUST 11-13: 25th Annual New Jersey QSO Party. Spon- sored by the Englewood Amateur Radio Association, Inc. From 2000 UTC Saturday, August 11 to 0700 UTC Sunday, August 12 and 1300 UTC Sunday, August 12 to 0200 UTC Monday, August 13. Phone and CW considered same contest. Sug- gested frequencies: 1810, 3535, 3900, 7035, 7135, 7235, 14035, 14280, 21100, 21355, 28100, 28610, 50-50.5, and 144-146. Exchange: QSO number, RST, and QTH (ARRL sec- tion or country). New Jersey stations send county for their QTH. Logs and comments to: Englewood Amateur Radio Association, Inc., PO Box 528, Englewood, NJ 07631-0528. Include a #10 SASE for results.

SEPTEMBER 9-15: The Southern Counties Amateur Radio Association (SCARA) is planning to have a special events sta- tion during the Miss America Pageant. Check September Ham Radio for details.



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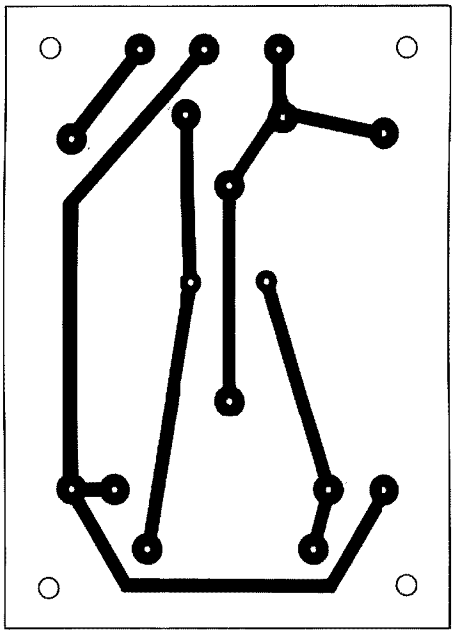

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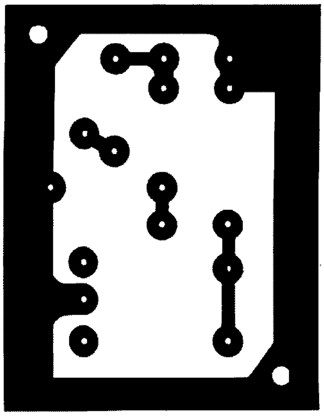

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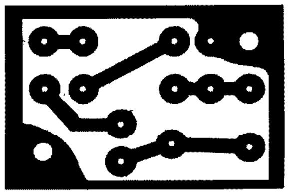

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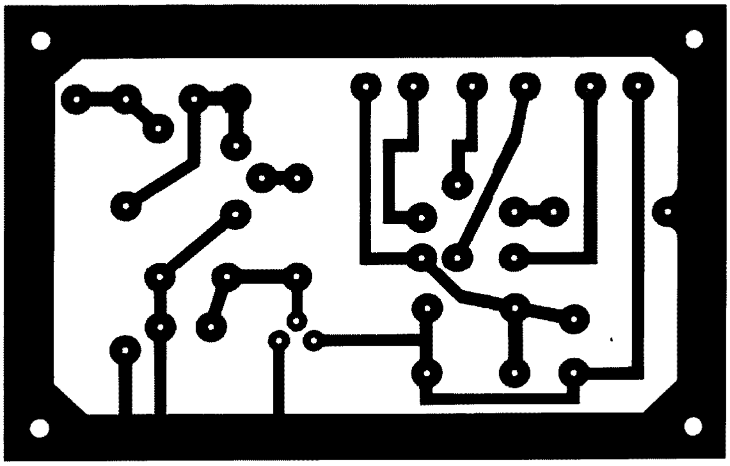

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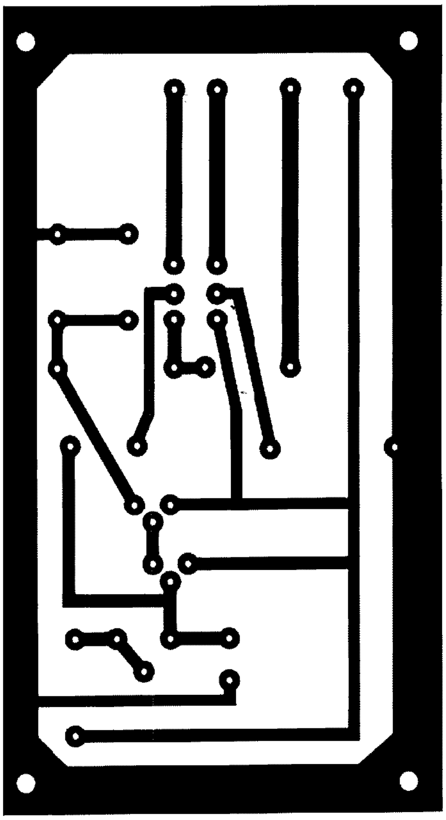

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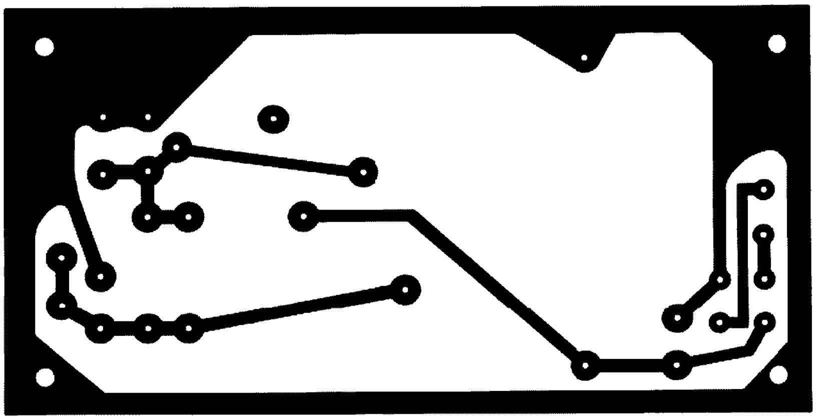
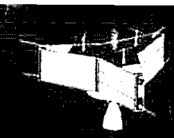


F 26 VDC regulated power supply

SEPTEMBER 1984 / $2.50



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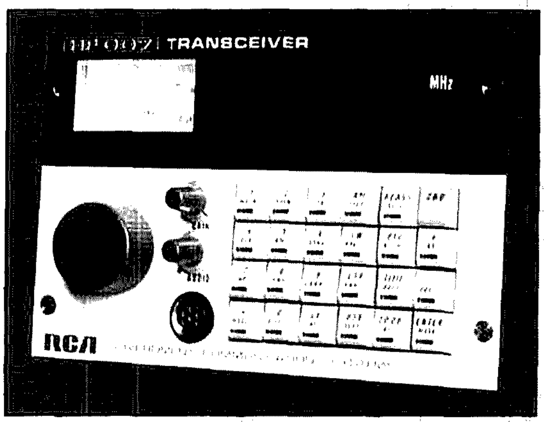
FIRST HT-TO-HT
SATELLITE QSO

ham radio magazine



CELEBRATING 70 YEARS OF PROGRESS:
THE DEVELOPMENT OF SSB

hr
focus on communications technology

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SEPTEMBER 1984

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ham radio

magazine

contents

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REFLECTIONS

taking care of business

There are some controversial arguments to be made in behalf of software piracy. These are some of the arguments we've heard:

> *Software is overpriced. Piracy gets it into the hands of users at prices they can afford.*
>
> *Piracy encourages the development of new and improved programs to meet the rising expectations of a sophisticated (and possibly saturated) market.*
>
> *As pirated programs circulate, product visibility and name recognition are enhanced. This helps manufacturers market current and future products.*
>
> *Software piracy is no more than a logical extension of the problem of trying to protect copyrights in the audio and video industries. As such, it can't be stopped.*

One industry maverick even claims that by providing real competition to software manufacturers, piracy actually helps keep software prices *down.*[1]

It occurred to us that similar, even identical, arguments could be made in defense of burglary. Burglary, after all, provides part-time employment to some (the burglars) and full-time employment to many (the police). It supports free enterprise and R&D — in security systems, insurance, medical care, and penology. It also gets consumer goods into the hands of users at a price they can afford.

We understand that nobody cares more about copyright law than those whose livelihood is directly affected by copyright infringement. Publishers — of books, music, software, and magazines — care very much, because illegal reproduction of thier product affects their ability to continue producing that product. Is it unreasonable to ask or expect the general public to care? Perhaps. But we think it's not unreasonable to ask the Amateur Radio fraternity to care.

We asked one software publisher how many hours his firm had invested in the development of a program for CW/RTTY/ASCII transmission and reception. "Thousands," he said, and explained that it had taken the firm's chief programmer *six months* of full-time effort to develop the program. Add to this the cost of support services and debugging; of developing associated hardware; of writing clear, comprehensive manuals; and of advertising and overhead. Suddenly the $29.95 or even $99.95 price tag on a package of Amateur Radio software becomes a little more understandable. The software pirate, on the other hand, needs nothing more than a computer, a disk drive, a supply of disks, and access to a photocopier for duplicating *original documentation* to go into business.

Some consumers don't know that unwrapping a package of software constitutes acceptance of a *limited* license to use the program inside. We'd like to think that anything we buy is ours, to do with what we will. After all, we can buy a Hershey® chocolate bar and share it with a friend, but we can't take it home, duplicate the recipe, pour our own chocolate bars, stamp them with Hershey's good name, and sell them as our own. The license to use isn't a license to steal.

The fact is that none of the manufacturers who serve the Amateur Radio market exclusively, or nearly so, are listed among the Fortune 500. Some are virtual mom-and-pop operations. Others are small groups of entrepreneurs. A few have grown and diversified. Many have prospered. But there's not one among them who can afford to keep fighting software piracy without sacrificing continued investment in new products.

Every dollar the Amateur Radio software industry spends fighting piracy is a dollar that can't be invested in R&D. If we want products that can keep pace with our rapidly expanding interest and needs, we'd best put our money where it can help make the difference — not in some pirate's pocket.

Dorothy Rosa, KA1LBO
Assistant Editor

1. "Can Software Makers Win the War Against Piracy?" *Business Week*, April 30, 1984.

presstop de W9JUV

ADDITIONAL PHONE BAND FREQUENCIES WILL BECOME AVAILABLE on September 1, following the Commission's adoption of a further Report and Order on PR Docket 82-83 July 18. The new phone subbands will be exactly as proposed in the FCC's Further Notice of Proposed Rule Making that came out over a year ago: for 75 meters, 3750-3775 kHz, Extra; 3775-3850 kHz, Extra/Advanced; 3850-4000 kHz, Extra/Advanced/General. On 15, 21200-21225 kHz, Extra; 21225-21300 kHz, Extra/Advanced; 21300-21450 kHz, General as well. For 10, the phone band bottom edge was moved down to 28300 kHz for all three license classes.

40-Meter Phone Was Unchanged For Continental U.S. Amateurs, but Amateurs in Alaska and the Pacific were given 7075-7100 kHz to make their operation compatible with Amateurs in ITU Region 3 and avoid Region 3 short wave broadcasters on 40 meters' high end.

THE VULNERABILITY OF 220 MHZ TO TAKEOVER BY ANOTHER SERVICE was sharply underscored by FCC Private Radio Bureau Chief Bob Foosaner at the July 21, NYC ARRL National Convention's FCC Forum. His remarks and responses to questions left listeners little doubt that Amateur use of part of the band, probably the bottom 2 MHz, could end in the very near future.

220 MHz For Commercial Users Was Also Addressed in a petition filed with the FCC in late June by the Land Mobile Communications Council. In its petition the LMCC, a trade group of 2-way users and manufacturers, reviewed various spectrum options for the ever-increasing needs of the land mobile service. 220-225 MHz was cited due to its "relatively low" Amateur activity along with its availability to land mobile in ITU Region 2.

THE ARRL WAS APPOINTED A VEC IN ALL 13 DISTRICTS in a ceremony at the ARRL National Convention in New York July 21. The proposal accepted by the FCC was actually the ARRL's second; its earlier proposal had raised questions about the "Chinese Wall" between League publishing efforts and its VEC administration, but the new one (submitted only July 20!) cited organizational changes which seem to adequately isolate the two League activities from each other. Now that exam fees are permitted, ARRL-sponsored hamfest exams are planned in September. However, League exams for individuals still won't be available until November.

Exam Fees Up To $4 Were Authorized By The FCC July 12, when the Commissioners acted on Docket 84-265. Justification of VEC fee scheules will be required; in the beginning, VEC's must estimate the cost of their programs, then set their fees by dividing that cost by the projected number of examinees. They're then required to maintain proper expense records; if the fees later turn out to be higher than actual costs, the "excess" must be returned by adjusting fees for later applicants downward an appropriate amount. VEs as well as VECs will be permitted to recover their costs, with division to be settled between them. Fees may be collected after August 31, but there is no requirement that fees must be charged

DeVry Is Considering Applying For VEC Status In Seven More Districts, through DeVry Amateur Radio Society members on its campuses in those districts. The school has been well pleased with the results of the Society's VEC program on its Chicago campus, and feels the program would be a worthwhile addition for the other DeVry schools as well. The additional campuses are in Los Angeles, Phoenix, Kansas City (Missouri), Atlanta, Columbus (Ohio), Dallas, and Woodridge, New Jersey. The DeVry campuses in these cities would be able to offer regularly scheduled walk-in exams, as Chicago already does, plus support and even testing facilities for VECs already in place if they wished to use them. DeVry-administered VE groups are now giving exams in various parts of Illinois and Indiana, and DeVry gave its first Advanced Class exam in Chicago on July 24.

VEC District 13, The Pacific, Should Have A Resident VEC very shortly thanks to the Koolau (Hawaii) Amateur Radio Club. Now that the ARRL has also been certified, all 13 districts have at least one resident VEC.

New Identifiers For Amateurs Upgrading In The VEC Program have been adopted by the Commission for August 31 implementation. These are /KT for Technician, /AG for General, /AA for Advanced, and /AE for Extra. The unique session identifiers are being dropped.

PERMISSION TO BROADCAST ON 40 METERS FROM GUAM is being sought by Trans-World Radio Pacific, a religious broadcaster. Guam is in ITU Region 3, which allocates 7100-7300 kHz to broadcasters, but the FCC has not previously licensed 40-meter broadcasting in areas under its jurisdiction. Comments on the proposal, RM-2959, are due in mid-September.

Clarification Of Broadcasters' Use Of Amateur And CB Communications is being sought in RM-28-30, proposed by the Commission July 12. The proposal would relax present Part 73 and 97 requirements that Amateurs and CBers give prior permission before their communications could be rebroadcast. Reporters, however, would still not be permitted to operate an Amateur station for news gathering. The Comment period ended in mid-August.

TWO AMATEUR RADIO OPERATIONS FROM SPACE SEEM LIKELY IN 1985, as W5LFL has been named a crew member for Space Shuttle Mission 51H, scheduled for next November. WØORE is already scheduled for Mission 51F in March, and indications are that NASA is reacting favorably to the joint ARRL/AMSAT proposal that Amateurs on future shuttles be permitted to operate.

AN AMATEUR RADIO AUXILIARY HAS BEEN FORMED BY THE FCC'S FIELD OFFICE BUREAU, in conjunction with the ARRL's Communications Department. Assistance in monitoring and rules enforcement will be the major focus of the new Auxiliary, which, like the VEC program, is a result of Senator Goldwater's bill. Like the VEC program, coordination with the FCC will be through regional or national Amateur organizations. ARRL is the first such group to sign up. Individual Amateurs interested in joining the Auxiliary should contact John Lindholm at ARRL or Elliott Ours at the FCC, 1919 M Street N.W., Room 744, Washington, D.C. 20554 (202) 632-7090.

comments

ground rod resistance

In the interest of brevity, Mason Logan, K4MT, intentionally omitted a detailed step in the derivation of the equation shown in Appendix A of "Ground Rod Resistance" (July, 1984, page 95). For readers who wish to follow the progress of this derivation without interruption, K4MT has prepared an addendum to his article, available upon request from *ham radio*. Enclose a business-sized SASE. Editor

modular receiver

Dear HR:

Since publication of my article, "A Modular Two-Band Receiver," in July, 1983, I have received and answered well over 150 inquiries.

It seems there is quite a bit of interest in construction articles about communication receivers. I have had letters from all states and from India, South Africa, Australia, New Zealand, and several other countries.

Over the past year, however, some minor errors in the article have come to light. These errors have been corrected in an errata sheet available from *ham radio* (send SASE).

James J. Forkin, WA3TFS
Pittsburgh, Pennsylvania

seaworthy sealant

Dear HR:

I enjoyed Bill Orr's column in the January issue in which he explained the realities of coaxial cable construction. I was horrified, though, when he suggested use of RTV sealant for weatherproofing coaxial connectors. If you've ever tried to remove that stuff years later, you know what I mean.

May I suggest an improved method that we use aboard ship? First wrap the connector with self-adhering rubber tape (3M or similar), extending the tape approximately 2 inches past the connector end. Next, wrap in a similar fashion with regular plastic electrical tape. Then apply several coats of Scotchkote™ (3M) weatherproofing.

This technique has proven very effective in this most demanding environment. Years later, it is a pleasure to remove this "cocoon" with one slice of a knife, and find a shiny connector as good as it was on the day of installation!

Scott W. Barber, WA2DRL
USS Trenton (LPD-14)

ground plane loop

Dear HR:

I would like to call attention to a conflict in the details given in Bill Orr's February, 1983, column of the ground-plane loop antenna. Both the drawing (**fig. 1**) and the text indicate that the semicircle is to be 0.2 wavelength long. In contrast, **table 1** shows the dimension to be 0.1 wavelength. The article in *CQ-DL* doesn't clarify this either.

C.T. Atherton, WD6DUD
Bell, California

Mr. Atherton is correct. This same conflict appears in all three references (ham radio, Radio Communication and CQ-DL). Do any readers know what the correct tap distance should be?

software piracy

Dear HR:

There is a problem in the Amateur Radio fraternity: software piracy. Whether by ignorance or simple disregard for the law, many Amateurs are stealing copyrighted programs. Most do not consider their theft a crime or a serious problem, but unless this practice is discontinued Amateur Radio will suffer.

With the influx of computers into the hobby a degree of software piracy was inevitable. Unfortunately the problem has become a blemish on Amateur Radio. Thousands of dollars have been spent in litigation involving software piracy outside the hobby, and I had hoped Amateur Radio would police the problem internally and not require legal action. Sadly this is not the case.

I recently confronted two hamfest exhibitors who were selling copies of a Kantronics program. These people were copying and selling our programs to any Amateur willing to pay the price. I bought one of the programs for evidence and informed the seller that legal action would be taken. This seller was not a ham, but those buying the program were. We have several other examples of programs copied and sold.

There are a few simple steps we can all take as those interested in seeing the problem solved.

1. Never buy copied software.
2. Report pirates to the software manufacturer.
3. Don't allow illegal sales at your local hamfest.

Kantronics plans to prosecute those who steal our programs, as we have in the past. But without the assistance of the entire Amateur community the manufacturers will not be able to stop pirates from stealing their profits. If manufacturers are not able to sell enough products to make a profit other new and improved programs will not be written. Don't let the greed of a few deny the hobby of future expansion. Let's throw the bad apples out before they ruin the whole barrel.

Mike Forsyth
Marketing Director
Kantronics, Inc.

An Ohio-based dealer charged with selling pirated Amateur Radio and other types of software at a Michigan hamfest has been fined $2000 for violation of United States copyright law. In its July 17 decision, the federal court in Toledo awarded damages to Kantronics, who initiated the suit, and issued a permanent injunction against further production and distribution of the illegal software.

— Editor

the development of Amateur SSB: a brief history

70 years of progress marks this familiar mode

Single-sideband radio telephone transmission in the Amateur bands — even well into the VHF region — is taken for granted these days. But it wasn't always so. Many Amateurs remember the days when amplitude modulation with full carrier (AM) reigned supreme. (For some it still does — Editor.)

While the era of Amateur SSB is generally considered to have begun with an article published in 1948 by Arthur Nichols[1], much work was done well before; basic groundwork, in both theory and in hardware, was completed as early as 1933 by the ARRL and other groups. Unfortunately, much of this work was considered too complex for the Amateur community, and was consequently not published.

early history

The earliest written analysis of an amplitude modulated signal consisting of a carrier and upper and lower sidebands separated by the modulating frequency was made by Carl R. Englund,in a paper dated August 19, 1914.[2] Englund who worked for the Bell System, recorded his analysis in his engineering notebook, but no record of what — if any — use was made of this information survives today. There is evidence that others were aware of the existence of sidebands, but Englund's analysis is apparently the earliest record to survive. In 1915, for example, H.D. Oliver, a telephone company engineer working transatlantic radio telephone at NAA, the then-new Navy radio station at Arlington, Virginia, considered the use of SSB for solving the problem of communications. Oliver proposed to tune the antenna to eliminate the carrier and one sideband, which would have been entirely practical because the transatlantic tests were made at 5000 meters (60 kHz), where antenna *Q*s are necessarily very high. But because the tests were carried out using AM, nothing appears to have come of this idea.

The fact that SSB, with or without a carrier, was considered as a means of transmitting information implies not only knowledge of the existence of sidebands, but also of the fact that information is contained therein.

No attempt was made to develop SSB for radio communication purposes at this time, probably because of the complexity of the receiver that would be required. State-of-the-art receivers in those days consisted of several stages of RF amplification — a detector followed by several stages of audio amplification. Likewise, filter design, too, was primitive by today's standards.

first SSB transmitted on a wire

The only organization to demonstrate interest in SSB was the Bell System, which developed the technique for long distance wire-line telephony. The first commercial wire carrier system using SSB was placed in service in 1918. SSB techniques were used exclusively for long distance telephone circuits until relatively recently, when they were replaced by pulse-code modulation (PCM).

The first application of SSB to radio was made in 1922, again by the Bell System. Bell set up an SSB transmitter, operating on 57 kHz, at Rocky Point, Long Island, and the British Post Office established a receiving station at New South Gate, England, near London. One-way communication was established in January 1923, proving the feasibility of SSB for transatlantic communication. The return message from England came by transatlantic telegraph cable. When a transatlantic radio telephone circuit was put into commercial operation on long wave (57 kHz) in 1927, SSB was used. As the traffic increased and additional channels were needed, these were operated in the

By John J. Nagle, K4KJ, 12330 Lawyers Road, Herndon, Virginia 22071

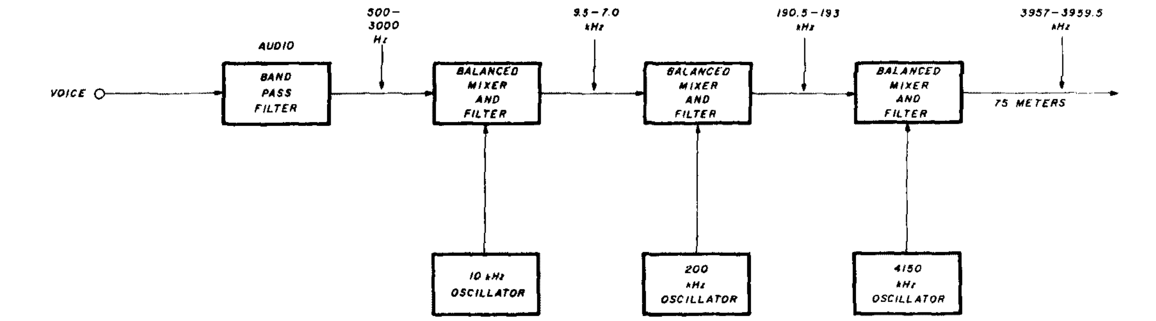

fig. 1. Block diagram of 75-meter SSB transmitter developed by Robert Moore circa 1933.

short-wave band (HF, in today's terminology). Interestingly, short-wave circuits used AM *because* SSB techniques for HF were not perfected until 1936.

amateur SSB

As has often been the case in the history of Amateur Radio, one or two hams or rather groups of hams developed techniques for Amateur communications that closely paralleled work going on in industry. In the early 1930's, while Bell was adapting the SSB techniques used on wire-line to HF radio for transatlantic telephone circuits, two Amateur groups were working on Amateur SSB.

The first record of the development of an SSB transmitter for Amateur use that I have been able to find appeared in a series of three articles by Robert M. Moore, then W6DEI, in a magazine called *R/9*; a block diagram of Moore's transmitter is shown is **fig. 1.**[3] (Old-timers will remember *R/9* as a first-rate technical magazine. In fact, one of the issues describing Moore's SSB transmitter also features an article on how to build a parabolic antenna — interesting reading even today! The name *R/9* was lost when the monthly publication merged with *Radio* in January, 1936, after which *Radio* itself became a first-rate technical magazine.) Moore acknowledges having drawn on the published data of the Bell System "for a portion of the material used" in his series.

The second group working on SSB was an ARRL team led by James J. Lamb, then Technical Editor of *QST*. At the Board of Directors' meeting held on May 12, 1933, Bernard J. Fuld, W2BEG, director of the Hudson Division, moved:

> *"That the technical staff of **QST** is instructed to investigate the feasibility, and, if feasible, is instructed to undertake the development at reasonable prices, of apparatus and methods of single-sideband and carrierless 'phone transmission'."*

The motion passed unanimously.[4]

As a result of this motion, K.B. Warner, W1EH, who was secretary of the League as well as Editor-in-Chief and Business Manager of *QST*, sent a memorandum to James Lamb; because of its historical interest and the insight it provides into the personalities of the people involved, it is reproduced in **fig. 2**.

The memorandum was followed by a short ARRL inter-office memo (**fig.** 3) adding directional antennas to SSB and encouraging a "serious attempt to accomplish something." The list of names at the top of the memo is worth reviewing; how many of these people can you remember?

Lamb and his associates did go into a huddle, as requested; Lamb wrote a 12-page report, which he forwarded to Warner on September 25, 1933. While the report is too long to reproduce here, an excerpt (**fig. 4**) is included because A.L. Budlong (then ARRL Communications Manager) wrote some comments on this part of Lamb's report. **Fig. 5** shows Budlong's comments on Lamb's SSB report.

Although Lamb's report was never published in *QST* as originally written, a later, revised version was published in the October, 1935 *QST*.[5] This article was not listed in either the table of contents for that issue or in the annual index published in the December 1935 *QST*.

The decision of ARRL management not to pursue development of SSB as a means of reducing phone band interference was apparently based on the assumption that SSB equipment was too complicated for the average Amateur in 1933-34.

At that time, it was the policy of League officials, as well as the *QST* editorial staff, to do everything possible to encourage more people to become licensed Amateurs. This was believed necessary for the preservation of Amateur frequency allocations in the short-wave bands that were (and still are) essential for its continuation. In 1933-34 there were only about 20,000 licensed Amateurs — a small force in comparison to the highly organized, well-funded

JJL:

By vote of the Board of instructors, we are instructed to investigate the feasibility of applying to amateur operation single-side-band and carrierless phone transmission, and if we find it feasibile we are further instructed to produce cheapest-possible apparatus to accomplish it.

I now place this matter in your hands. Suggest you and your cohorts do a bit of reading up on it and then go into a huddle. I shall then need a written report of what you find and what you propose to do about it. I believe that that report will either state that there is ND and give definite reasons why not, or will find ND except in a limited extent in which you propose certain investigation and further report, or (theoretically at least) will find that there is good possibility of accomplishing something and that lab work is going to be undertaken along such-and-such lines, with a further report to come.

It is my instant impression that there is ND. I think that that will be the offhand opinion of all of you. I ask you-all, however, to put on your amateur thinking-caps and look at this from the traditional amateur point of view that there is always a new way to skin a cat. Without any doubt at all any such development would be of tremendous importance -- as an amateur achievement, as a commercial development of vast worth, as an immense practical aid in congested amateur phone territory. To be acceptable, your report will have to show to the Board why it is not practicable to build such amateur phone stations, e.g., because it would have to have a filter that cannot be produced for less than $3000.

I do not know whether an autodyne receiver is capable of supplying the carrier for suppressed-carrier transmission or not. If it is not, and if it is not possible to use any local oscillator easily to supply the missing carrier, I should think that method of small interest to us.

I also wish to suggest your consideration of partial-sideband ~~xxxxxx~~ suppression if you do not find it practicable to attempt full suppression of one sideband. I mean, a method using a filter so as to cut off all frequencies above the value necessary for reasonably good speech, thus suppression unnecessarily high voice frequencies, harmonics, etc. I believe that a filter to do that can be built for a very small fraction of the cost of one necessary for eliminating one sideband neatly.

....

At this stage in my note-writing you and Hebie have used up my remaining time but I have just discovered that I have, compact in one booklet, all the studies made of this thing by the COIR and I am going to conclude by handing you this publication of Opinions, calling your attention to pp. 59 to 74 relating to this subject. The booklet is to be returned to me, please.

HW?

KBW
KBW 6/6/33

PS— Ur reports, of course, are to be transmitted by me to the Board

fig. 2. K.B. Warner, League secretary and Editor-in-Chief and Business Manager of *QST*, sent this memo to *QST* Technical Editor James Lamb.

American Radio Relay League
OFFICE MEMO

Beaudin	Hebert	Mr. Maxim
Beakley	Houghton	Mr. Stewart
Budlong	Houldson	Mr. Segal
Chamberlain	Hull	
De Soto	Lamb	
Grammer	Rodimon	
Handy	Scanlan	Mr.________

Please note and return to me.
Returned noted; thanks.
Referred to you for necessary action.
For your files.
Please give me information on marked portion.
Please note and file.
What do you think about it?
For your information.
When finished with your part, please pass on to ________________

I have just discovered that it was Fuld's intention to include in that motion re sideband suppression, etc., the subject of directional antennas. That is a separate and large order, but one that you might take a look at in your conference.

I concluded my note of yesterday without saying (as I meant to, that I want you fellows to make a serious attempt to accomplish something along the desired lines if it is feasible.

fig. 3. Brief inter-office memo urged League staffers to make a "serious attempt to accomplish something."

entrepreneurs and industrialists representing commercial communications and international broadcasting, who wanted to take over the Amateur frequency assignments for their own use.

The League felt that the best way to increase the number of licensed operators was to describe equipment that was "sure-fire" in all League publications; that is, equipment that was easy to build with simple hand tools and certain to work without elaborate test equipment or much experience on the part of the builder. There was concern that if a newcomer started out in Amateur Radio by building the latest equipment — an SSB rig for example — he would probably be unable to make it work, and would become discouraged and leave Amateur Radio.

Because Amateur Radio is flourishing today, we can only speculate about what the outcome might have been if the Amateur community had actively pursued SSB in the early 1930s. If the Bell System, with all its resources, which were substantial even then, was not able to develop practical SSB equipment for HF radio until 1936, it seems likely that it would have been all but impossible for individual Amateurs to have done it in 1933 or 1934. (Even today, most Amateurs prefer to buy, rather than *build*, their SSB equipment.)

The League did stick to their "keep it simple" philosophy, despite the frustration of some technical people at ARRL headquarters who would have preferred to have been able to continue their pioneering developmental work in Amateur SSB for professional reasons.

Even though the ARRL did not pursue the develop-

-3-

The first of these, suppressed-carrier double-side-band transmission, would eliminate the steady-type interference resulting from the heterodyning of undesired carriers with the carrier of a desired station and would reduce distortion from selective fading, but would require the same frequency band-width as would normal (carrier and double-side-band) transmission. Simple suppressed-carrier transmission also might have an economic advantage in that the full capability of a linear r.f. stage could be utilized for the intelligence-carrying side-band *power to the exclusion of the carrier power* that normally is transmitted. In a 100% sinusoidally modulated wave, 2/3 of the total power is represented by the carrier and 1/3 by the side-bands.[2,3] For instance, whereas a truly linear amplifier of 300-watt maximum capability would have but 100-watt side-band output with a normal 100% modulated wave, with the carrier suppressed the side-band power with the same amplifier theoretically could be as high as 300 watts, representing three times the intelligence-carrying power of the carrier-and-double-side-band capability. From the amateur point of view this advantage is more apparent than real, however, because a given r.f. amplifier when operating Class-B and [redacted] handling only side-band power has negligible advantage in side-band output over the same amplifier when operating Class-C with plate modulation in the system of carrier-and-double-side-band transmission generally used by amateurs. ~~Difficulty with the simple suppressed-carrier system that at the receiving end will be discussed under "Reception"~~.

The second system, transmission of the carrier and one side-band, with the other side-band suppressed, is technically less feasible than

-4-

suppression of the carrier and elimination of one side-band because elimination of one side band without affecting the carrier would be more difficult than suppression of the carrier with elimination of one side-band. This system hardly merits consideration. Although it would seem to offer advantages in that the frequency band required would be lessened by the one set of side-band components eliminated, it would offer no opportunity of realizing greater intelligence-carrying output from a transmitter of given peak-power rating than is obtained with carrier and double-side-band transmission. The gain would be negligible in view of the additional technical complexities involved. The heterodyne interference problem would be in no wise lessened.

The third system, ~~Of the three types mentioned,~~ single-side-band transmission with the carrier suppressed, appears to be not only the most economical in frequency-band requirement and in utilization of transmitter capability, but also to hold less technical complication in reception than the first-mentioned type (double-side-band), and still less technical complication in transmission than the second type (carrier and one side-band). It also eliminates most of the distortion from selective fading. The present objective must be primarily the minimizing of the interference problem, so far as that may be possible, to the end that greater [redacted] utilization of the available amateur 'phone bands may be hoped for. Hence, only single-side-band suppressed-carrier transmission warrants our serious consideration. This view is reinforced by the receiving considerations to be discussed later.

Carrier Suppression

Two means of suppressing the carrier-frequency component are possible. One would be a filter having sharp attenuation at the carrier frequency and passing either one set or both sets

fig. 4. Excerpt from Lamb's 12-page report, sent to Warner on September 25, 1933.

Bottom p.3, top p.4. I do not agree. The sole aim of Board's hope is to increase effective width of fone bands by changing amateur radio to a system that requires fewer kc for a transmitter. Jim departs from the point when he damns this system by saying it offers no opportunity to increase output from given tubes. Nobody cares. Nobody cares even if output is decreased -- if stations now occupy only half as much room.

Considering difficulty of replacing carrier, and of tuning at receiver with suppressed-carrier method, I hardly can embrace it. I am most reluctant to accept Jim's top of page 4 that it is too difficult to suppress just one sideband. He there makes bold statements with no supporting references. His language looks more like prejudice than logic that the reader embraces because it is self-warranting.

At this moment I've read only first 3½ pages, but I here make note that I'm pretty sure that the only feasible idea is suppression of one sideband.

Bottom of page 4 needs a statement to this effect: "All right, then, let's now have a look at methods that give this form of transmission. The first thing to consider is suppressing the carrier...." Hiatus now.

Having read it,

It is too goddamnably impartial. Too deep a subject for me to do my own deciding, I'm ready to accept the author's views, but I don't even know what he thinks. With almost no effort, he could make me think either way. It needs just a bit of bias here, since it really is headed somewhere -- to prove something or other,-- isn't it? "Further in proof of the futility of aspiring to so and so..... "Another example of the impracticability...."

(Imagine my urging bias here, condemning it in 2d paragraph)

fig. 5. A.L. Budlong, ARRL Communications Manager, added comments to this copy of Lamb's report.

ment of SSB in the 1930s, individual Amateurs did. James Lamb continued to think about it and proposed some interesting ideas — revolutionary in those days — including the transceiver concept of using the same oscillator for transmitting and receiving that is the backbone of all SSB transceivers today.

Fig. 6 shows an SSB generator using Lamb's crystal filter circuit. This sketch is dated June 18, 1933 — over 50 years ago. Notice the signatures of witnesses: F. Cheyney Beeklye, *QST* Advertising Manager, and Ross A. Hull, Associate Editor.

Another interesting and novel (in 1934) circuit for a modulator is shown in **fig. 7**. Here, two tubes are connected with the anodes in push-pull and their control grids in parallel. RF excitation is applied to the parallel connected control grids. This represents the carrier frequency, which is eliminated in the push-pull anode circuit. The audio information is connected, push-pull, into the suppressor grids. This circuit has the advantage that the RF and audio are kept separated. (This disclosure was witnessed by Ross Hull.)

Probably Lamb's most dramatic development was the SSB transceiver sketched in **fig. 8**. This is basically the same block diagram as the SSB transceivers that began to appear on the market *in the early 1970s*. One slight difference may be interesting, however. Lamb perceived a receiver tuning problem with SSB. With AM, the practice was to tune to maximize the carrier. With SSB, this was no longer possible, because there is no carrier. To solve the receiver tuning problem, Lamb envisioned transmitting a 1 kHz tone; the receiver would be tuned until the 1 kHz tone was *actually* 1 kHz. When SSB systems came into actual use, the tuning problem was solved by Amateurs in a much simpler manner. Amateurs simply tuned the receiver until the voice sounded the most "natural."

Unfortunately, these developments in the use of SSB for Amateur communications were not publicized.

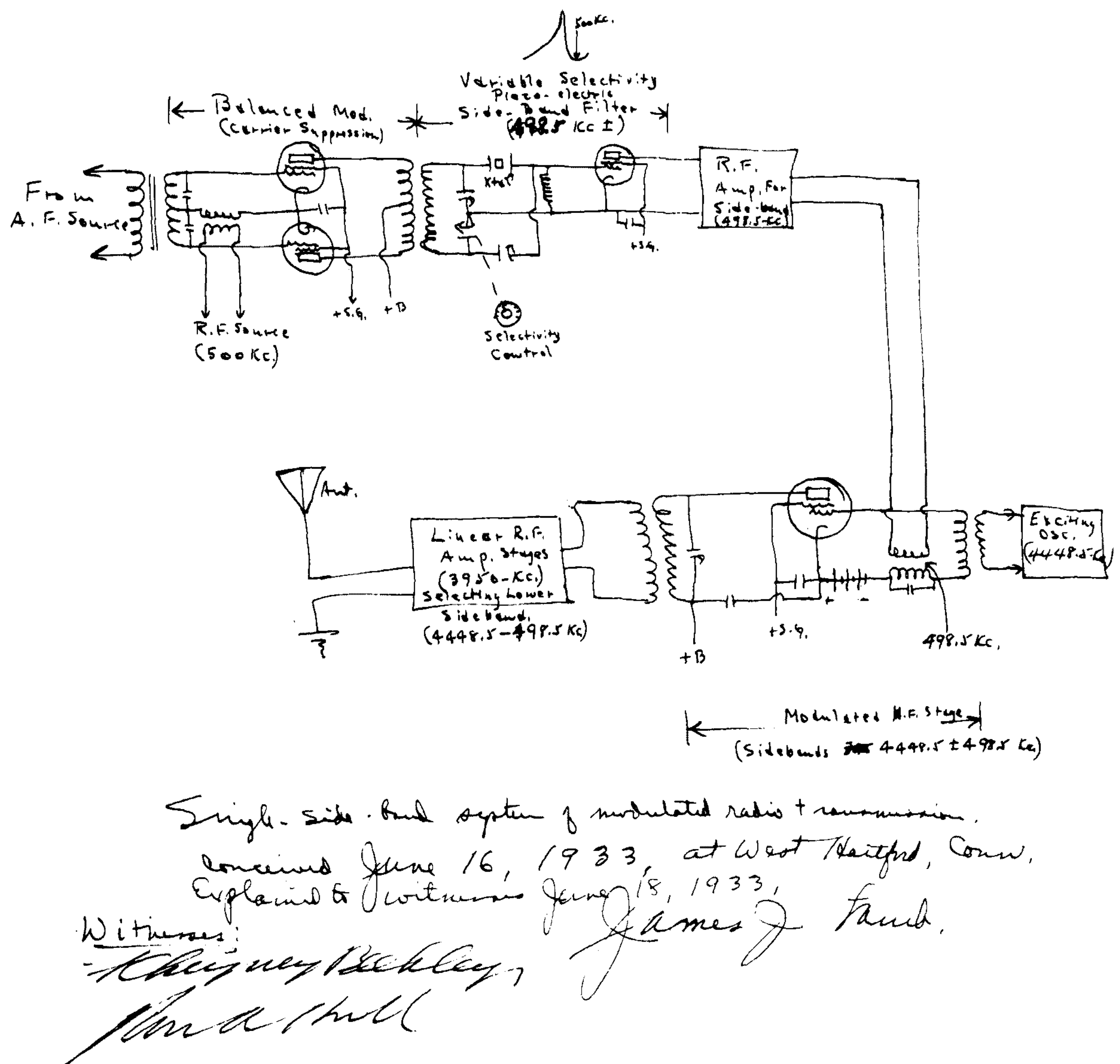

fig. 6. Lamb's early design of an SSB transmitter using his crystal filter.

Amateur SSB — phase two

The second phase in the development of Amateur SSB began during and immediately after World War II. O.G. Villard, then W6QYT, had become interested in SSB during the war. Although he was aware of Moore's work in 1934, he was intimidated by the filters required by Moore's rig. This apparent handicap led him to develop a phasing technique to generate SSB. As often happens to inventors, Villard discovered that the phasing technique had already been devised, and had been patented in the late 1920s.

The phasing technique offered the advantage of eliminating the need for elaborate filters and reducing the number of frequency translations required. Filter technology was still in its infancy, and few Amateurs had access to the test equipment necessary for verifying a filter design. The biggest component problem lay in the design and construction of the required inductors, with adequate *Q*s in the required frequency range.

As most readers will recall, with the phasing method of SSB generation, two components of the carrier 90 degrees apart in phase are developed; this was not difficult to do in the early 1940s. Similarly, two components of the audio voltage — also 90 degrees apart in phase — are developed. This posed much more of a problem for Villard, who attempted to use one phase

unmodified and to develop a wide band, a 10:1 frequency range, and an audio channel shifted 90 degrees in phase. This proved to be a real problem because of the relative bandwidth required.

This problem was not solved until R.B. Dome[6] showed that it was not necessary to actually shift the audio signal 90 degrees, but only to develop two audio voltages that were 90 degrees apart and then describe networks to do this. Fortunately, Dome's networks required only resistors and capacitors — no inductors. With this development, Villard was able to proceed with the actual construction of an SSB transmitter.

At the time, Villard was teaching electrical engineering at Stanford University. He and some of his students constructed an SSB transmitter using the phasing method and operated it on 20 meters from the Stanford Amateur Radio Club's station W6YX.[7] Surprisingly, many Amateurs were able to copy SSB with their AM receivers; their reports were extremely positive. Villard's transmitter used four 813 tubes in the output stage and was capable of much more power output than the filter-type transmitters of the time. The SSB signals were generated at the operating frequency (20-meter band) so that no frequency translations were required.

About ten days after Villard's group began transmitting, another SSB signal appeared on 20 meters. Arthur Nichols, then WØTQK, had built an SSB rig using the filter approach in only about five days, after hearing W6YX's SSB signal. This was a remarkable accomplishment, especially for that time; even today, most Amateurs would have difficulty building an SSB rig in five days.

filter approach to SSB

The January 1948 issue to *QST* featured an article by Nichols describing his SSB transmitter. Two related articles also appeared in this issue: one by Byron

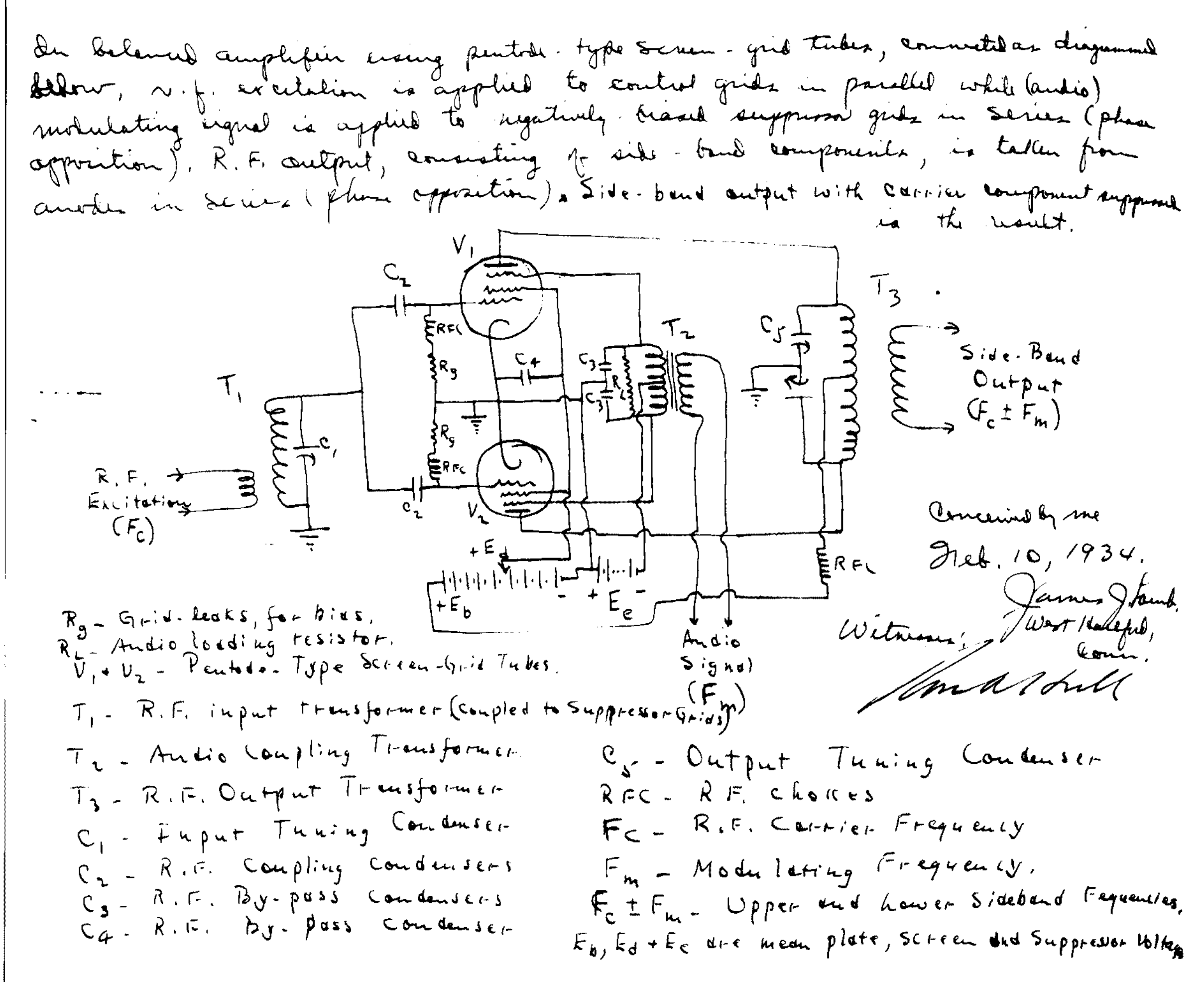

fig. 7. Modulator circuit separates RF and audio.

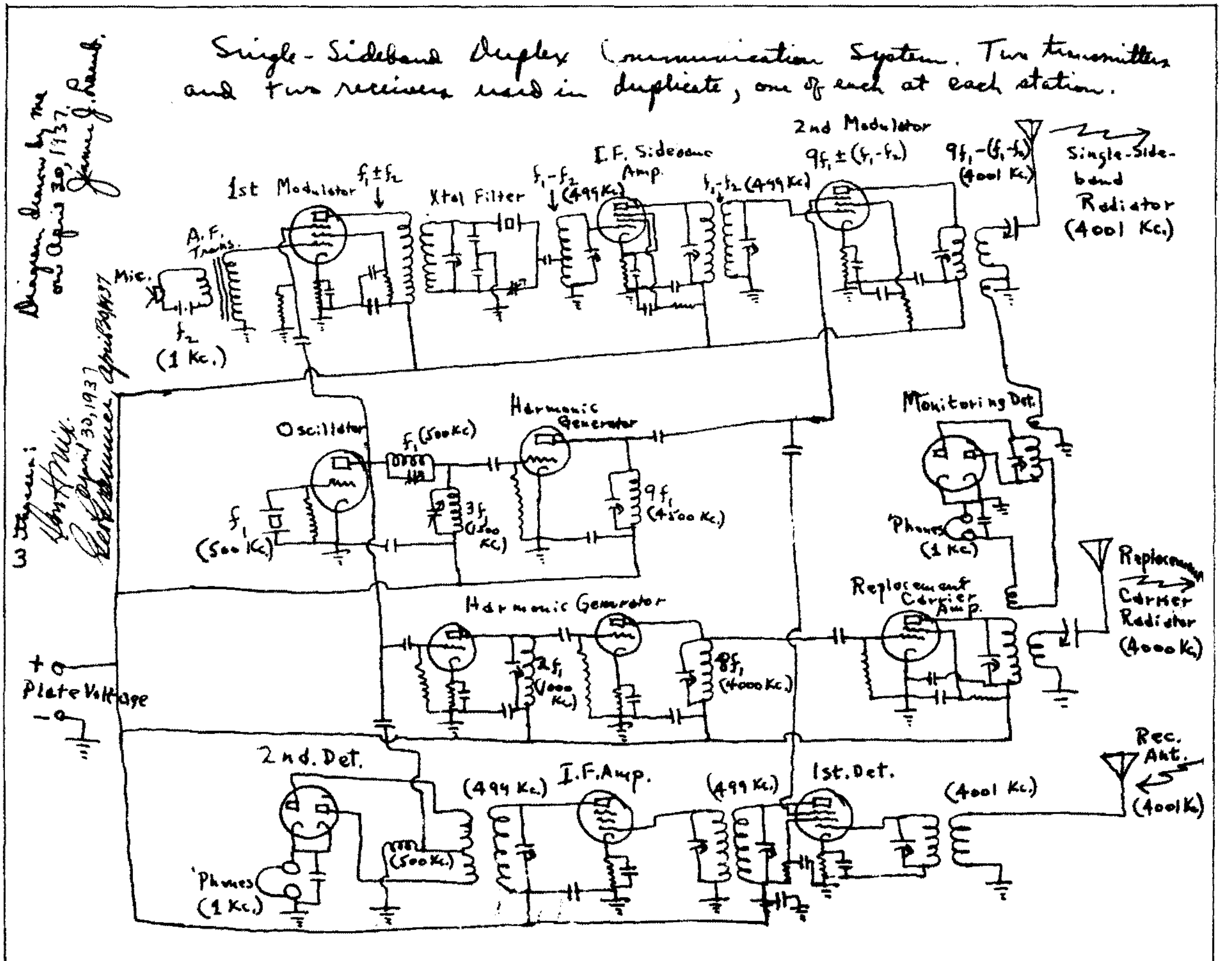

fig. 8. Lamb's SSB transceiver.

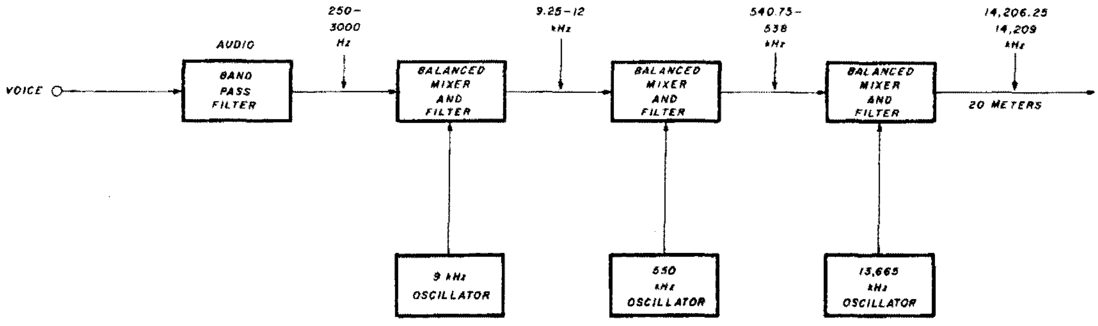

fig. 9. Block diagram of 20-meter SSB transmitter developed by Arthur Nichols circa 1948.

Goodman, W1DX,[8] who explained the theoretical aspects of SSB, and a second by Villard,[9] in which he described the early SSB on-the-air tests from W6YX.

A block diagram of Nichol's 1948 transmitter — similar to Moore's 1933 rig — is shown in **fig. 9.** Nichols used a 10-kHz filter to eliminate the carrier and undesired sideband with two frequency translations to 20 meters. Moore used two translations to 75

meters. From a cursory examination, it appears that Nichol's and Moore's filters used Bell System technology; both appear to have been based on telephone carrier filters, with Moore using the next lower carrier channel slot than Nichols. Nichols obtained his filter from the late Fred Berry, formerly W0MNN.

By this time SSB enjoyed the active support of the League and others: additional theoretical material and improved techniques and equipment appeared in rapid order. In fairness, it must also be pointed out that the level of technical sophistication of the Amateur community had been greatly raised by the rapid development of electronics during World War II and by the large number of people introduced to the new technology both at home and in the field. This — coupled with the increase in the ranks of Amateur Radio operators and the fact that work done by Amateur Radio operators was more widely recognized — improved the political posture of Amateur Radio in its struggle against commercial and international broadcast interests in the battle for frequency allocation.

final comments

Arthur Nichols designed and built his SSB transmitter independently of other workers in the field. First licensed in 1931, he has always been a home-brew type, and is currently W6EVL in Fallsbrook, California.

In addition to the SSB transmitter Nichols used himself, he built two more transmitters which he sold to the National Company of Malden, Massachusetts. At the time, National was one of the leading manufacturers of Amateur equipment. I assume Nichols' transmitters were to be prototypes for a National SSB transmitter; if so, National could have had an early lead in supplying SSB equipment to the Amateur community. But National was unfortunately having financial problems and chose not to take advantage of the opportunity.

Nichols' interest in SSB came about naturally. His father was the late Dr. H.W. Nichols, an engineer with the Bell System who worked on the development of SSB radio equipment for transatlantic radio telephony service. Dr. Nichols was sent to England in 1923 to work with the British Post Office on arrangements to receive the Bell System transmissions mentioned earlier. While in England, he presented a paper before the British Institution of Electrical Engineers on the transatlantic radio telephone problem.[10] The elder Nichols died in 1925 and did not live to see the long-term results of his work.

I have not been able to locate Robert M. Moore, formerly W6DEI, and I believe, with regret, that he may be a silent key. I would be happy to hear from anyone who knows (or knew) Moore, and particularly from anyone familiar with his interest in SSB.

acknowledgements

My thanks to Professor O.G. Villard, W6YX, Arthur H. Nichols, W6EVL, and James Millen, W1HRX, who reviewed an early draft of this manuscript and suggested changes, many of which have been incorporated.

My special thanks to James Lamb, who made his engineering notebook available to me, thus arousing my interest in the history of SSB.

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3. Robert M. Moore, formerly W6DEI, "Single Sideband Transmission," *R/9*: Part I, July/August, 1933, page 7; Part II, December, 1933, page 18; Part III, January, 1934, page 25.
4. Minutes of Annual Meeting, (May 22, 1933), Board of Directors, American Radio Relay League, *QST*, Volume XVII, No. 7, July, 1933, page 24.
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9. O.G. Villard, Jr., "Single-Sideband Operating Tests," *QST*, Volume XXXIII, No. 1, January, 1948, page 16.
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ham radio

an audio automatic gain control

Circuit converts 50 dB input dynamic range to constant output

In a recent *ham radio* column,[1] Bill Orr commented — correctly, I believe — on the relative lack of use, in Amateur Radio, of state-of-the-art developments in audio processing technology. Perhaps this is so because commercial audio processors are very expensive, or because the broadcast quality frequency response and distortion characteristics of commercial equipment represent a higher degree of precision than is really necessary in Amateur Radio applications — a kind of technological "overkill."

This article describes an automatic gain control (AGC) amplifier that can be used in several Amateur Radio applications. Simple and inexpensive to build, it will maintain, at a maximum gain, a constant output level within ±2 dB over an input level range of 50 dB.

As a broadcast engineer, I developed this circuit in hopes of eliminating the need for announcers to continually adjust levels for program material arriving by twisted pair. Amateur Radio applications could include autopatches, phone patches, and repeaters. (Imagine a repeater that is modulated to a constant level regardless of the received signal's deviation!) When used ahead of any transmitter, the AGC functions much like a compressor used in commercial broadcast operation.

circuit description

The AGC schematic is illustrated in **fig.** 1. An audio signal applied to U1, an MC3340P, is passed through to the 741 operational amplifier, U2. After being amplified, the output signal of U2 is sampled and applied to a negative voltage doubler/rectifier circuit composed of diodes CR1 and CR2 along with capacitor C1. The resulting negative voltage is used as a control voltage that is applied to the gate of the 2N5485 JFET Q1. Capacitor C2 and resistor R2 form a smoothing filter for the rectified audio control voltage.

The JFET is connected from pin 2 of the MC3340P to ground through a 1-kilohm resistor. As the voltage applied to the gate of the JFET becomes more negative in magnitude, the channel resistance of the JFET increases causing the JFET to operate as a voltage controlled resistor.

The MC3340P audio attenuator is the heart of the AGC. It is capable of 13 dB gain or nearly −80 dB of attenuation depending on the external resistance placed between pin 2 and ground. An increase of resistance decreases the gain achieved through the MC3340P. The circuit gain is not entirely a linear function of the external resistance but approximates such behavior over a good portion of the gain/attenuation range.[2]

An input signal applied to the AGC input will cause the gate voltage of the JFET to become proportionally negative. As a result the JFET increases the resistance from pin 2 to ground of the MC3340P causing a reduction in gain. In this way the AGC output is held at a nearly constant level.

Because a finite time is needed to generate the feedback to control the AGC gain, an abrupt change from soft to loud at the input will cause a short overshoot or "pop" sound. Capacitor C3 with resistor R3 form a low-pass filter in the feedback circuit of the 741 operational amplifier. This low-pass action minimizes the overshoot.

adjustments

There are three possible adjustments to the AGC. They are dynamic range, attack time, and recovery time.

The dynamic range is adjustable by selecting the value of resistor R3 in the feedback circuit of the 741. A 100 kilohm value for R3 results in the maximum obtainable dynamic range, nearly 50 dB. Because this amount of gain will probably be too large for most Amateur Radio applications, R3 may be decreased to produce the dynamic range desired. If R3 is changed, C3 must also be changed. The product of R3 and C3 must remain the same constant value to preserve the low-pass filter characteristics of the stage.

The attack time is controlled by C1. The 0.33 μF value shown for C1 produces the fastest possible at-

By Lee Barrett, K7NM, 525 North 2150 West, West Point, Utah 84015

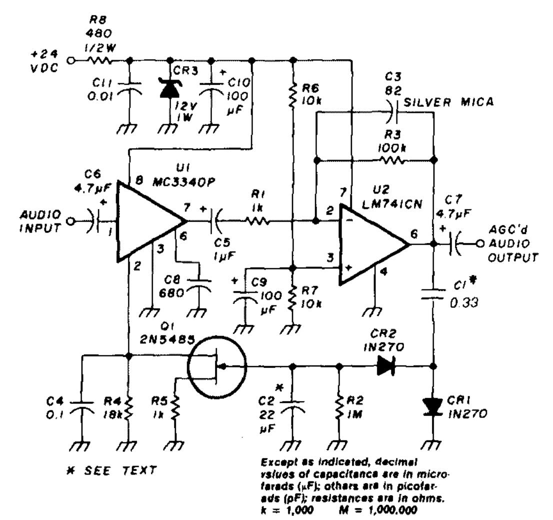

fig. 1. AGC schematic diagram.

parts list for fig. 1

item	*description*
C1	0.33 μF, 50 volt mylar
C2	22 μF, 25 volt radial electrolytic
C3	82 pF, 500 volt silver mica
C4	0.1 μF, 50 volt ceramic disc
C5	1 μF, 25 volt radial electrolytic
C6,C7	4.7 μF, 25 volt radial electrolytic
C8	680 pF, 500 volt silver mica
C9,C10	100 μF, 25 volt radial electrolytic
C11	0.01 μF, 50 volt ceramic disc
CR1,CR2	1N270 germanium diode
CR3	1N4742 12 volt, 1 watt zener diode
Q1	2N5485 N-channel JFET
R1,R5	1000 ohm
R2	1 megohm
R3	100,000 ohm
R4	18,000 ohm
R6,R7	10,000 ohm
R8	470 ohm, 1/2 watt
U1	MC3340P voltage controlled attenuator
U2	LM741CN operational amplifier
miscellaneous	PC board, IC sockets, optional transformers (see text), solder, wire, case, builder's choice of connectors, 24 volt or alternative power supply (see text)

Notes:
All resistors are 1/4 watt, 5 percent unless otherwise noted.
A printed circuit board with parts kit ($26.00) or the printed circuit board alone ($15.00) is available from the author, Lee Barrett, K7NM, 525 North 2150 West, West Point, Utah 84015. Please add $2.00 shipping and handling.

table 1. AGC measured parameters with 0-8 bridging transformer input.

frequency response (−3 dB):
below AGC threshold (−40 dBm input): 200 Hz - 12 kHz
in AGC range (−20 dBm input): 40 Hz - 20 kHz
AGC threshold: −38 dBm
noise floor (input shorted): −42 dB maximum

input level (dBm)	output (0 dB = 2V p-p)	distortion (percentage)
10	2	2.0
−20	0	0.5
−36	−2	0.5

tack time. In no case should C1 exceed this value or a low frequency oscillation will occur.

Capacitor C2 is the main recovery time adjustment. I have found the best value for speech applications to be 22 μF. This value can be decreased or increased to provide faster or slower recovery times, respectively.

connections

The AGC may be fed either directly with an unbalanced input or through a transformer by a balanced audio source. In broadcast applications, I used high grade transformers such as the TRW 0-8 or 0-30. In Amateur Radio applications where speech is the rule less expensive Calectro transformers have been used successfully.

The AGC output is normally unbalanced. Loads as low as 600 ohms have been driven by the AGC although a 1 kilohm or higher resistance termination is desirable. Again, a transformer could be used to create a balanced output condition if desired.

In Amateur Radio applications a level potentiometer will probably need to be added across the output to act as a level adjustment. The relatively high level output of the AGC can then be reduced to the drive level required by the intended load.

conclusion

Table 1 lists the test results of measurements made on the AGC.

The original AGC was designed to plug into the Collins/Autogram IC series broadcast mixers and operated on 24 volts DC. In your application, however, the zener may be adjusted or eliminated along with the series resistor to operate the AGC on voltages down to about 8 volts.

references

1. William I. Orr, W6SAI, "Ham Radio Techniques — Ancient Modulation," *ham radio*, February, 1984, page 65.
2. *MC3340P Data Sheet No. DS9249R1*, fig. 4, copyright 1975, Motorola Semiconductors, Box 20912, Phoenix, Arizona 85036.

ham radio

work OSCAR 10 with your HT

Use a local gateway to work the bird

Soon, working intercontinental DX may be as simple as picking up your 2-meter HT. Using a local "gateway" facility* to relay signals to a satellite, thousands may soon sample the world of Amateur Radio space communications. How will this be possible? How can you participate?

AMSAT OSCAR 10 (AO-10) — the newest and most sophisticated Amateur Radio communications satellite ever — was launched in June, 1983, as a replacement for the ill-fated Phase IIIA. AO-10 (Phase IIIB prior to launch) has quickly established itself as the all-time DX champ of OSCARs; in the year or so since its launch, nearly 100 countries have become active on it. Intercontinental QSOs are now commonplace . . . in fact, some stations have worked over 90 countries already! Until the AO-10 was launched and operating, it took a moderately well-equipped VHF/UHF station to accomplish this; now intercontinental QSOs on modest equipment are commonplace.

AO-10 uses two linear transponders that receive inputs on one frequency and translate them downward to another band (see **fig. 1**). The transponders are functionally similar to a repeater with a wide frequency split. In the case of AO-10, a block of frequencies uplinked to it in the 435-MHz range is repeated on a correspondingly wide block of frequencies in the 145.8-MHz range; this is the Mode B transponder. A second, even broader transponder uses a 1269-MHz uplink and a 436-MHz downlink. (See **table 1** for the exact frequencies.)

*A buffer between you and the satellite which reduces your equipment requirements by interfacing your station with AO-10.

the gateway station

A gateway station provides many of the functions that an ordinary OSCAR earth station might perform. Virtually any station that can reach a gateway can get a taste of satellite activity. Here's how it works.

Let's assume the gateway station is associated with your local repeater and uses part of the repeater's equipment. In this case the gateway will, on *uplink*, take the audio feed from the repeater's FM receiver; drive a moderately powerful SSB transmitter uplink with the audio from the repeater; point the uplink antenna; and derive control signals (transmit/receive) from the repeater.

On *downlink*, the gateway will preamplify the SSB downlink signal from the satellite; feed the signal to an SSB receiver; take the audio from this receiver to the repeater FM transmitter audio input; point the downlink antenna; and derive control signals (transmit/receive) from the repeater.

The system illustrated in **fig. 2** can be considered as a standard OSCAR satellite station with multiple remote access.

As a broadband repeater, AO-10 takes a spectrum 150-kHz wide on the uplink, translates the frequency down, and repeats the signals with the same relative amplitude in the downlink spectrum. The Mode L transponder provides a greater capacity of 800 kHz of spectrum in which to work. That's more than all the spectrum in the 20 and 15 meter bands combined. A second type of gateway uses a linear transponder similar to the linear transponder used on the satellite, but with the frequency pairs reversed; in this manner several stations can simultaneously access AO-10 through this type of gateway. Both types of gateways have been tried successfully. WB3EYB in Harrisburg, Pennsylvania, has operated a gateway successfully in conjunction with a standard FM repeater. KE3D in

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table 1. AO-10 uplink/downlink frequencies.* (Note: These do not represent "channels" per se, but show the relationship between inputs and outputs. Coverage is continuous from band edge to band edge.)

uplink (MHz)		results in downlink (MHz)
mode B		
435.032		145.972
435.050		145.955
435.070		145.935
435.090		145.915
435.110		145.895
435.130		145.875
435.150		145.855
435.170		145.835
435.175		145.830
	general beacon	145.810
	engineering beacon	145.987
mode L		
1269.050		436.950
1269.100		436.900
1269.150		436.850
1269.200		436.800
1269.250		436.750
1269.300		436.700
1269.350		436.650
1269.400		436.600
1269.450		436.550
1269.500		436.500
1269.550		436.450
1269.600		436.400
1269.650		436.350
1269.700		436.300
1269.750		436.250
1269.800		436.200
1269.850		436.150
	general beacon	436.020
	engineering beacon	436.040

*exclusive of Doppler shift.

Boulder, Colorado, and DJ4ZC in Marburg, West Germany, are known to be working on the terrestrial, broadband linear transponder approach. The FM repeater approach is simplest and can be implemented quickly using existing equipment. The terrestrial linear transponder, on the other hand, requires special equipment and techniques.

There is nothing special about the choice of input/output frequencies for the station connecting to the actual gateway. Two-meter repeaters will work as well as 6-meter or 440-MHz versions. Suitable isolation techniques must be observed, however, because a strong 2-meter FM transmitter close to the gateway station 2-meter receiver is likely to be strongly affected (desensed) by the local FM signal. Remoting, cavity filtering, and other techniques are appropriate for this and similar situations where the repeater frequencies and the gateway frequencies are in the same band.

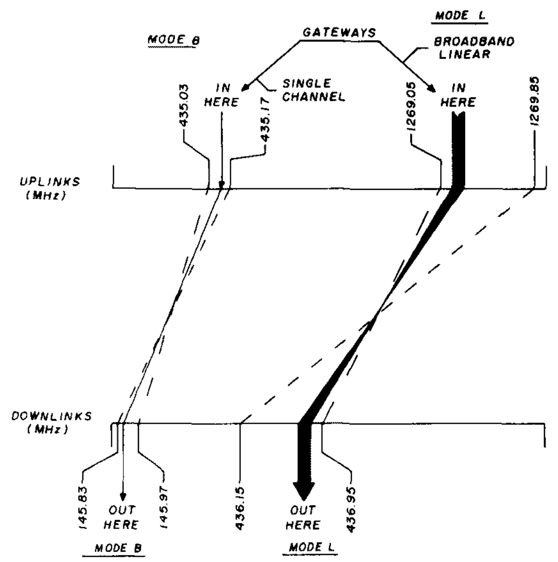

fig. 1. AO-10 modes and typical gateway operation.

operation

SSB and CW are the preferred modes of operation on AO-10. FM and full-carrier AM are discouraged because these continuous-power modes use precious solar-derived electrical power even in the absence of modulation; SSB does not. That's the reason FM inputs to the repeater are baseband converted to SSB via the audio circuitry. Under normal conditions, you should expect a received signal-to-noise ratio of 10 dB or more at the gateway. Under ideal, low-traffic conditions, the S/N ratio may approach 20 dB. The downlink power of the satellite transponder is shared among all the signals appearing in the uplink passband. A S/N ratio of 3 or 4 dB is normally sufficient for minimal copy. Thus, 10 to 18 dB will be heard, with intelligibility approaching good DX conditions on HF.

The gateway station operator needs to be especially sensitive to his or her responsibilities to both terrestrial communicators and to the manner in which the gateway "community" is introduced to the satellite community already using AO-10. The skill and courtesy of the operator are especially important when the uplinked spectrum is wider than normal, as with a terrestrial linear transponder 20 or 30 kHz wide. Future planning efforts will likely identify special gateway "zones" for uplinks to reduce the hazards of high traffic. AO-10 makes slightly more than two orbits per day. It travels in an elliptical orbit, which at times affords it a "view" of nearly one-third of the earth's surface. Its coverage area is extensive. Simply put, any station that can "see" the satellite can work any other that is simultaneously in view of AO-10, which will be in view for up to ten hours without interruption. Dur-

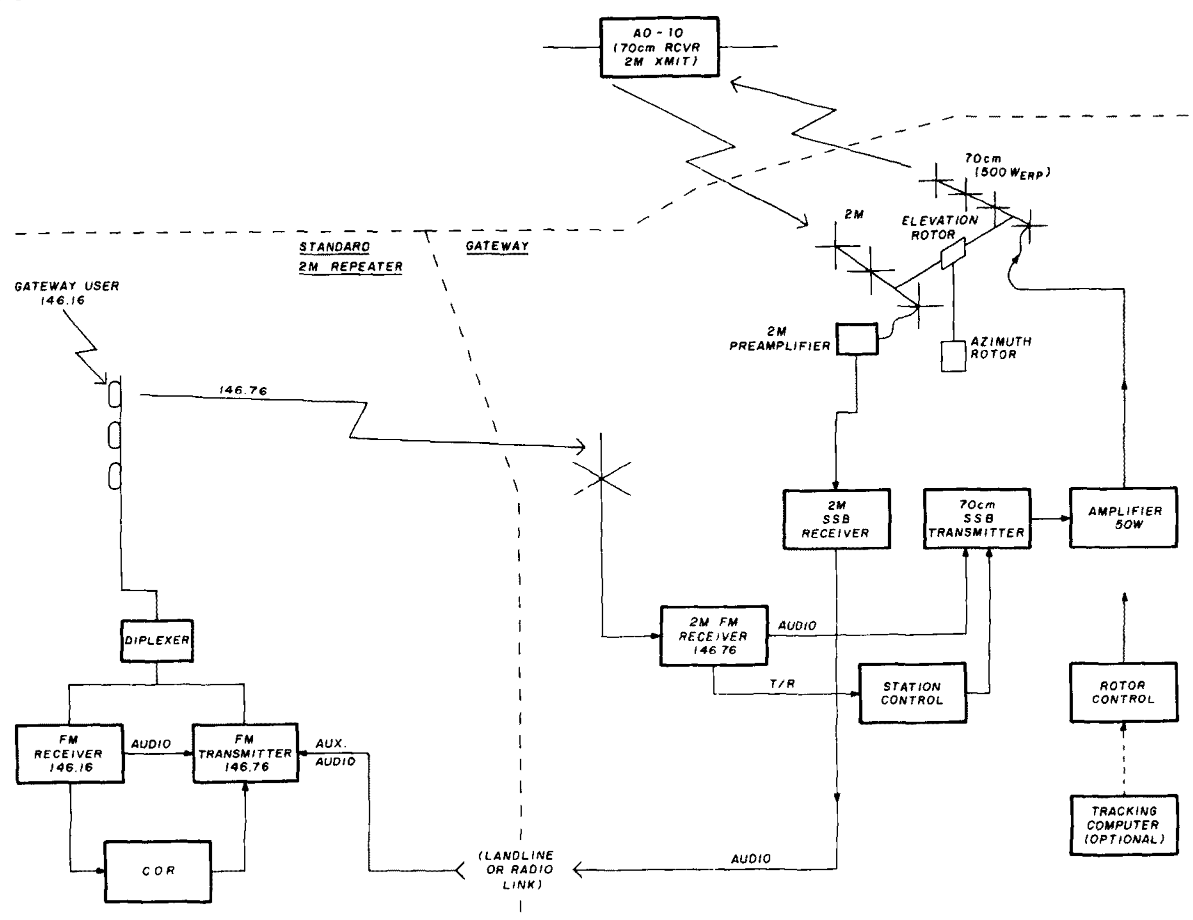

fig. 2. Typical narrowband (single channel) gateway plan using Mode B.

ing this period the satellite will move slowly across the field of view of the gateway station, whose antennas will track AO-10 either manually or under computer control. To the gateway user, however, operation will be "transparent"; that is, the gateway user does not need to know where the satellite is or the precise uplink/frequencies involved. The actual duration of gateway operation will be limited by a number of factors, including the gateway operator's schedule. For the present it appears that regulations require a full-time control operator be present; in any case, a fully automated gateway station is beyond the reach of all but a few.

Operation through a gateway is not intended, and certainly cannot, replace the fun and flexibility of establishing and operating your own autonomous satellite station. Assembling a station and learning the techniques required to be successful in satellite communications is not particularly difficult; it does, however, require some understanding of the basics. AMSAT, the organization that built and operates AO-10, invites the membership of interested individuals and organizations. One of AMSAT's main functions, aside from building and operating satellites, is instructing users and would-be users in their operation.

Whether working DX or just chewing the rag, nothing can beat satellite operating. Gateway access to AO-10 offers the newcomer an opportunity to "fly before you buy," in the sense that AO-10 operations can be sampled, without the expense of upgrading an existing station to full AO-10 capability. And for apartment dwellers who may experience difficulties in erecting suitable antennas (although many have done just that using a well-situated balcony, for example), using a local gateway may afford the best opportunity ever for working international DX from the comfort of your own easy chair, with your HT firmly in hand.

Further information and guidance on the use of gateways and membership in AMSAT is available from AMSAT, P.O. Box 27, Washington, D.C. 20044.

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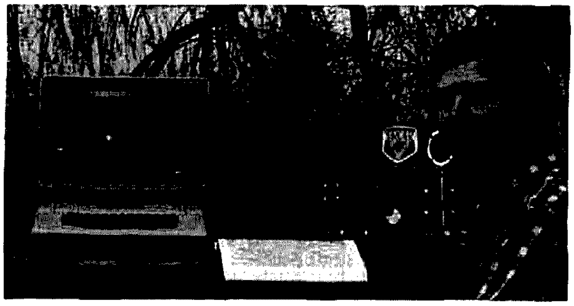
score first HT-to-HT QSO via OSCAR 10

de **WA2LQQ**



Mike Henderson, N6JFD, at controls of his gateway station in Camarillo, California

An important milestone in Amateur Radio was reached when two Amateurs in West Virginia and California became the first to QSO through an Amateur Radio satellite using 2-meter FM HT's. KB6DDQ in Camarillo, California, and KD8GL in Wheeling, West Virginia, established contact at 1458 UTC on May 28, 1984.

The historic event was facilitated by "gateway" stations that connected local terrestial repeater systems to the high-flying AMSAT OSCAR 10 (AO-10) satellite. Though CW QSO's in which one of the stations keyed the transmit switch of an HT as a crude key have been reported, it is believed that the May 28 QSO was the first in which *both* participants used HT's.

Operating from West Virginia, KD8GL used his HT to talk through the Triple States Radio Amateur Club Repeater, KD8GL/R. Signals from the repeater were picked up by the local gateway station, WB8ZTV, which converted from FM to SSB and to the 435 MHz OSCAR 10 uplink frequency. Signals were then beamed by WB8ZTV to OSCAR 10, high over the Western Hemisphere.

In Los Angeles, meanwhile, gateway station N6JFD tuned to the AO-10 downlink frequency and converted the SSB back to FM, retransmitting the signals to the WA6OBT repeater, to which KB6DQ was tuned. The return path to Wheeling mirrored this path.

The repeaters and the gateway stations operated full duplex, and the QSO's were two-way. According to monitors, signals in both directions were excellent.

The W7LWE repeater/gateway in Lake Havasu City, Arizona, joined in later, making this the first three-way gateway operation and effectively linking Amateurs in three states via OSCAR 10's trunking capabilities.

Also participating in the May 28 linkups were KR3V, K8AN, K2QWD, N6IAW, W7MCF, and others.

Just one day before the 2-meter HT contact, the Wheeling gateway was linked for three continuous hours to the expert satellite station ZL1AOX in Christchurch, New Zealand, thereby confirming the long-duration coverage expected from OSCAR 10.

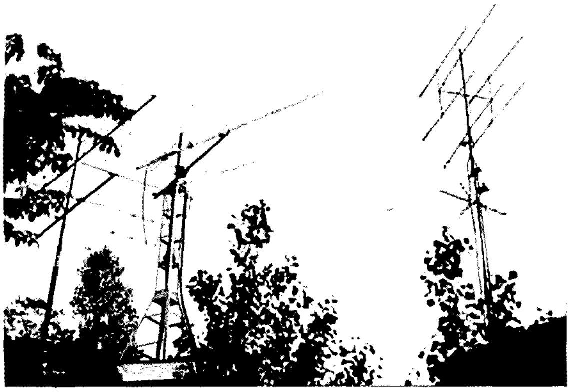
AMSAT suggests that using gateway interconnects may be the best way to demonstrate the capabilities of OSCAR 10 to prospective users, at no cost to them.

Don Knollinger, WB8ZTV, at controls of his gateway station in Moundsville, West Virginia.

Karen Henderson, KB6DDQ, operated one end of the link with only her HT from Camarillo, California.

For a free information kit, send an SASE to AMSAT, Department GW, P.O. Box 27, Washington, D.C. 20044.

Photos courtesy of Asterisk Design.



Uplink/downlink at WB8ZTV uses array at center; 70 cm up, 2 meters down.

KB6DDQ *tuned* to the WA6OBT/R repeater in Thousand Oaks, California, shown here with owner/operator Larry King.

Personally, I think that nothing can beat the flexibility of having your own OSCAR station, but for those just starting out, this seems to be a good way to taste the wine before you buy the bottle!

ham radio

the FM advantage

An Amateur's view of frequency-modulation theory

Do you ever wonder why you don't hear any ignition or static noise on your 2-meter FM transceiver but at the same time, often hear nothing but static on 40 meters? Or why the local FM broadcast station always sounds so much better than its AM counterpart? *Well, the answer lies in the theory* of modulation, and in the receivers that detect this modulation.

Although VHF-FM has become the most popular mode on the 2-meter band, few Amateurs really understand what goes on inside their "black boxes." This article should help to explain some of the apparent mysteries of FM.

historical development

Although the theory of frequency modulation had been explored many years earlier, it wasn't until about 1935 that a demonstration by Major Edwin Armstrong showed the advantages of FM over conventional amplitude modulation. These tests were conducted in the Yonkers, New York, area from Amateur Radio station W2AG, using VHF equipment.

Although World War II disrupted the development of commercial FM transmission, FM had, by that time, established a strong foothold in the communications field and was used during the war on frequencies higher than 30 MHz. The years following the war saw the growth of the commercial FM-broadcast industry with *its "wideband" high-fidelity music, and programming.* The middle 1950s brought about narrowband FM (NBFM) transmission, particularly on the high end of 10 meters. In the 1960s, channel agreements were negotiated for the 2-meter band, and a whole new era began — channelized 2-meter FM. This concept has spread to all of the Amateur VHF bands.

But why is FM better than AM? Let's look at some basic ideas about modulation.

modulation theory

Modulation can be broadly defined as the transmission of data on an electromagnetic wave. For common telegraphy (A1 transmission or CW) the presence or absence of a signal with respect to time comprises the modulation. Conventional amplitude modulation imposes a modulating signal upon an RF carrier. The resultant summation of these two signals is the variance in amplitude of the RF carrier at a rate of change equal to that of the modulating signal. That is, the amplitude of the carrier physically changes from some minimum level to some maximum level in proportion with the amplitude of the modulating signal. The frequency of the RF carrier remains the same. In an FM system, the modulating signal does not vary the amplitude of the carrier. Instead, the frequency of the *carrier* changes. This is the basic difference between AM and FM transmission, (see **fig. 1**).

Now that we know what is being changed by the modulating signal in an AM or FM system, let's see why it makes a difference. Major Armstrong correctly theorized that manmade and atmospheric static, or "noise," was amplitude-specific in nature. A lightning stroke generates RF signals at many frequencies (typically the entire LF through VHF spectrum) which all have different random amplitudes. The result is that we hear a loud, scratchy sound on conventional AM broadcast receivers, and on most AM, CW, or SSB receivers as well, no matter how good they are! The reason for this is that the AM receiver has no means of discriminating between the desired transmitted AM

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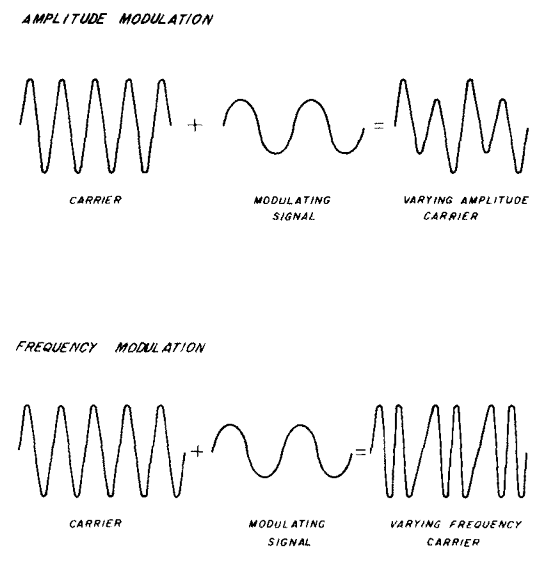

fig. 1. Basic differences between AM and FM transmission.

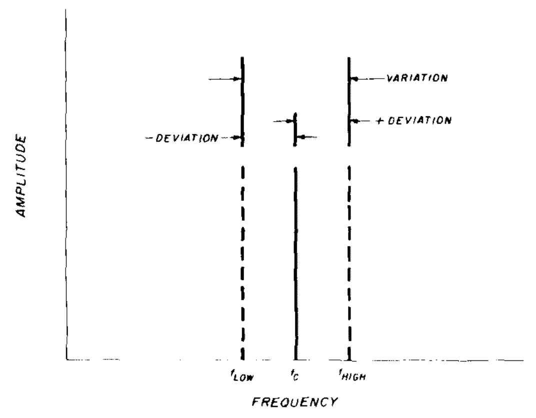

fig. 2. Relationship between carrier center frequency, deviation, and variation.

signal and the undesired AM static crash. The receiver is amplifying and detecting both signals simultaneously (as it was designed to do) and putting them both in your ear!

Consequently, amplitude-specific noise is difficult to detect with an FM system because there is no amplitude detector in an FM receiver! An FM receiver is looking for changes in frequency, not amplitude; it is for this reason that there is a qualitative improvement over an equivalent AM system with respect to static and noise.

terms

With an FM transmitter, three terms must be addressed before we can proceed with this discussion: deviation, modulation index, and deviation ratio.

Deviation is defined as the amount of frequency change of the carrier when modulated by a signal of a unique frequency — that is, if we were to apply a 1-volt peak-to-peak (e.g., +0.5 Vp and −0.5 Vp) signal of 700 Hz, it would cause the carrier to change frequency to plus and minus some specified amount at a rate equal to the modulating frequency, which is 700 Hz. This "specified amount" of frequency change is called the deviation. It is "specified" because the amplitude of the modulating signal is what specifies it; for example, a 2-volt p-p 700-Hz modulating signal creates more frequency deviation than a 1-volt p-p signal. (The deviation is adjustable by the gain of the modulating circuit.) From 0 to +0.5 volt, the carrier will deviate in the "plus-frequency" direction; from 0 to −0.5 volt, the carrier will deviate in the "minus-frequency" direction. Therefore, the total change of frequency is 2 times the deviation (see **fig. 2**).

Now that we are deviating the carrier, let's talk about *modulation index*. As we've just seen, the deviation is related to the amplitude of the modulating signal. But what about the modulating frequency? It would appear to have some effect — and it does. The modulation index is defined as:

$$mod\ index = \frac{deviation}{modulating\ frequency}$$

In effect, then, the modulation index is a ratio, or a pure number — i.e., the units of the numerator and the denominator cancel each other. It can be thought of as simply a parameter that describes the operation of the system.

The *deviation ratio* is similar to the modulation index:

$$deviation\ ratio = \frac{maximum\ frequency\ deviation}{highest\ modulating\ frequency}$$

One may consider the deviation ratio as a "maximized" modulation index. For example, a 1-volt p-p, 1 kHz modulating signal may deviate the transmitter plus and minus 25 kHz. Then:

$$mod\ index = \frac{25\ kHz}{1\ kHz} = 25$$

But what about a 1-volt p-p, 3 kHz signal?

$$mod\ index = \frac{25\ kHz}{3\ kHz} = 8.33$$

The modulation index is not necessarily the same under all conditions. However, if we define the 3-kHz tone as the highest modulating frequency that the transmitter will "see," and state that the 1-volt p-p amplitude will give the greatest frequency deviation

(25 kHz), then we define a system parameter or constant. For this sample, then:

$$deviation\ ratio = \frac{25\ kHz}{3\ kHz} = 8.33$$

Thus, the fine line between deviation ratio and modulation index has been established. But why is this important? Because the theory of FM improvement is based upon the modulation index (and deviation ratio) of the transmitted signal.

receiver noise

When there is no signal present at the input of a receiver, we hear "background noise" consisting of both ambient thermal noise generated by the movement of electrons in molecules of matter and noise generated within the components of the receiver. The magnitude of the noise may be calculated in terms of units of power — *i.e.*, watts per Hertz. Specifically, the ambient noise power in any 1 Hz of spectrum can be calculated by the equation:

$$noise\ power\ P_n = kT\ (in\ watts/Hz)$$

where K is Boltzman's constant $\approx 1.3803 \times 10^{-23}$ Joules/degrees Kelvin, and T is the absolute temperature (in degrees Kelvin) of the ambient surroundings. (Room temperature is about 290 degrees Kelvin.)

This formula can be expanded to determine the noise power of a specific range of frequencies by simply multiplying the power for 1 Hz by the total bandwidth to be analyzed:

$$noise\ power\ P_n = kTB\ (in\ watts)$$

where B, usually the receiver's IF selectivity value, is the bandwidth of spectrum under consideration in Hertz.

For a 2-meter receiver with 10 kHz of IF passband, the theoretical noise power is:

$$P_n = 1.3803 \times 10^{-23} \times 290° \times 10{,}000\ Hz$$

$$= 4.0029 \times 10^{-17}\ watts$$

Most communications engineers prefer to speak in terms of dB. This calculated power level will now be referenced to 1 watt:

$$P_n = 10\ log \frac{4.0029 \times 10^{-17}\ watts}{1\ watt}$$

$$= 163.97\ dBW$$

This number is the theoretical natural noise power level in 10 kHz of the 2-meter Amateur band at about 62 degrees F (290 degrees Kelvin).

If we had a perfect receiver whose components did not generate noise, then the noise-power level at the receiver output speaker would be the same as the noise-power input at the antenna. As we know, this is not the case; typically, our receiver might have a noise figure (the amount of internally generated noise) of, let's say, + 10 dB. Then, with the noise power as calculated, the receiver would have a noise threshold of − 153.97 dBW. That is, any carrier that arrives at the antenna terminal of the receiver with a power level of less than − 153.97 dBW would be buried in the noise.

Continuing this example, if a carrier had a power level of − 143.97 dBW, it would be + 10 dB above the noise power level. Or, stated another way, the carrier-to-noise ratio is + 10 dB. This is expressed as follows:

$$\frac{C_I}{N_I} = \frac{carrier\ power\ at\ input}{noise\ power\ at\ input}$$

$$= \frac{-143.97\ dBW}{-153.97\ dBW} = +10\ dB$$

Remember, even though division is implied by C_I/N_I, because we are working in dB, the quantities are subtracted for division operations and added for multiplication operations. This + 10 dB carrier-to-noise power is what is at the receiver's input. Now, if we listen to the output of the receiver, we have a signal coming out of the receiver at some level. If we remove that signal by, for example, turning off the transmitter, we then hear the noise level. This output relationship is defined as the output signal power divided by the output noise power:

$$\frac{S_O}{N_O} = \frac{output\ signal\ power}{output\ noise\ power}$$

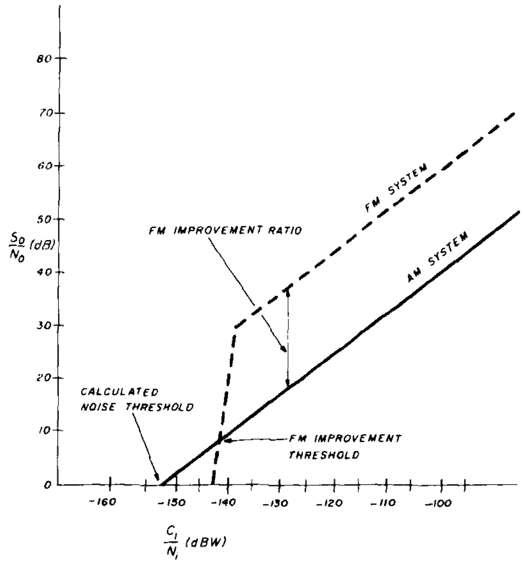

fig. 3. Typical input and output responses for AM and FM reception. Input numbers follow the example given in text.

This ratio, S_0/N_0, depends on the actual characteristics of the receiver. Obviously, a "quiet" receiver, for a given C_I/N_I, will have a relatively high corresponding S_0/N_0; if the receiver is of lesser quality, then N_o will be greater and the output S_o/N_o will consequently decrease.

AM and FM compared

This comparison between C_I/N_I and S_0/N_0 is often referred to as the signal-to-noise performance of a receiver, (see **fig. 3**).

AM noise performance shows an approximate 1:1 correlation — i.e., if C_I/N_I, increases by one unit, so will S_0/N_0. However, at a specific power level above the noise, typically about 10 dB above the noise threshold (− 153.97 dBW in our previous example), the FM system equals the AM response and starts shooting up more rapidly than the AM system. Therefore, for C_I/N_I, greater than about − 143 dBW as in our example, we see a higher corresponding S_0/N_0 with FM than with AM. The amount of improvement is called the FM-improvement ratio, and is directly related to the modulation index! The improvement ratio is:

$$\text{FM improvement ratio} = 3(\text{mod index})^2$$

Between − 153 dBW and approximately − 143 dBW, an AM system is superior to FM. However, for practical purposes, this improvement is not really noticeable in the noise.

NBFM compared

Let's look at our Amateur NBFM signals and see how they compare to wideband FM, SSB, and AM. **Fig. 4** illustrates this relationship between S_0/N_0 and input signal/noise level, S_I/N_I.* As can be seen from both **fig. 4** and the FM-improvement formula, the actual amount of improvement is greater with high modulation indices. Notice that wideband FM systems (such as are used by commercial FM broadcasting stations, and which are restricted to 75-kHz deviation by the FCC) provide the greatest amount of S_0/N_0 improvement. It does this, however, at the expense of weak signal detection; in other words, your FM stereo receiver may not be as sensitive as your AM/SSB communications receiver. However, for signals just above the minimum-detection threshold level, the output level shoots up very quickly, and the FM stereo receiver provides a better S_0/N_0 response to these weak signals than the AM/SSB equipment. The same is true for Amateur 2-meter NBFM transmissions (with 5-kHz deviation), with the exception that the sensitivity is improved at the expense of the amount of FM improvement. In practice, the NBFM improvement becomes apparent at such significantly low signal levels that we perceive FM detection to be invariably superior to AM. This is why FM has effectively replaced AM transmission on the 2-meter Amateur band even though AM is theoretically better at signal levels just slightly above the ambient noise level. Finally, we see that single-sideband suppressed-carrier systems offer a fixed improvement ratio over AM also. This can best be appreciated by realizing that the SSB voice channel is half of the bandwidth of a double-sideband AM system. Consequently, the ambient noise power in the SSB channel is less than in the AM voice channel. This results in a better S_I/N_I number and a correspondingly higher signal-to-noise output number.

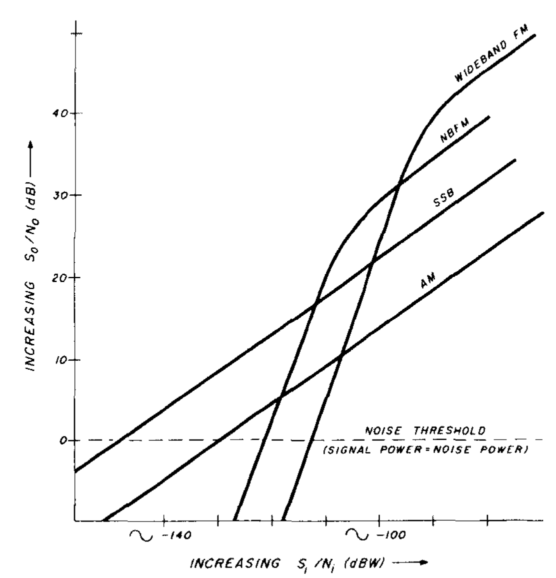

fig. 4. Typical receiver response curves for common modulation techniques. Power levels are approximate.

This is why FM 2-meter rigs seem to provide such clear voice reproduction. Next time you operate on FM, square your modulation index and multiply it by 3; you'll then know exactly how much improvement over AM is possible.

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*When referring to SSB modulation, the term Ci (carrier input level) is misleading since the carrier has been suppressed. Signal input level (Si) is more accurate, and for this analysis is essentially the same as Ci.

VHF/UHF WORLD

Joe Reisert W1JR

220-MHz EME requirements

The 220-225 MHz Amateur band has long suffered from the "I'll look into that band later" attitude. This wasn't helped by the fact that many services, especially the citizens' band radio service, had designs on this spectrum. Radars were also prevalent in some parts of the USA mainland, particularly near the coasts. Little or no commercial equipment was available for this band.

These negatives have all been turned around. The FCC has said "no" to the CB'ers and others seeking to claim this valuable frequency range; the radars have ceased operation; and commercial equipment is now available up to the Amateur Radio legal power limit.

Many VHFers have "done their thing" on 2 meters and are now looking for a new challenge. The 220-225 MHz band has much to offer them. In many ways it is like 2 meters without all the QRM! Equipment is very similar to 2 meters with smaller antennas a big plus, and local ambient noise and sky noise temperature are usually less than on 2 meters. Propagation is also similar, with meteor scatter and EME gaining increased popularity, especially since the introduction of the ARRL VUCC (VHF/UHF Century Club) award on January 1, 1983.

As this issue went to press the first intense sporadic E contact on 220-225 MHz had yet to occur.[1] The first few WAS certificates were recently awarded (in December, 1983), ironically several years after WAS was accomplished on a more difficult band, 70 cm (432 MHz). The first 220 MHz EME QSO didn't take place until 1970, well after it was accomplished on 2 meters, 70 cm, and 23 cm (1296 MHz).[2] This band has lagged behind the other bands, but that's changing fast.

With much of this month's issue dealing with specialized communications, I thought this would be a good time to review the EME (Earth-Moon-Earth) capabilities and operational requirements of this band. This information, especially in regard to equipment selection, can also be applied for those who just want to update their present gear for normal propagation modes.

minimum requirements for 220 MHz EME

The basic EME strategy should be to build a station that has the capability of hearing your own echoes, because this will allow you to evaluate your equipment and any other changes or improvements that you make.[3] Based on a 0-dB signal-to-noise ratio and no Faraday rotation (a random change in polarity that occurs when a VHF/UHF signal passes through the ionosphere), compliance with the following 220-MHz EME requirements will yield marginal echoes:

- path loss: 256 ± 1 dB
- minimum antenna gain: 22 dBi (approximately 20 dB over a dipole)
- maximum receiver noise figure: 2.0 dB (referenced to the antenna feed)
- minimum transmit power: 500 watts output (at the antenna feedpoint)
- receiver bandwidth: 500 Hertz maximum

Improvements above these minimum requirements such as increased antenna gain or output power will improve results accordingly.

antennas

As pointed out in reference 3, the most important part of an Amateur station is the antenna system because every dB of improved antenna performance yields a 2 dB system improvement: 1 dB on receive and 1 dB on transmit. This is especially important on EME, where antennas are large in comparison to those used on other modes of communications. Good clean patterns are also desirable with side, rear, and grating lobes (the lobes that result when two or more antennas are stacked) down at least 13-15 dB to get all the transmitted power aimed at the moon and to pick up the least amount of ambient or galactic noise (noise from outer space) on receive.

Yagi antennas, especially in this frequency range, are popular because they are easily constructed, easy to stack, and relatively small for the gain obtained. Hence, an array of high gain Yagis is presently the most common antenna used on 220-MHz EME.

The 4.2 wavelength NBS Yagi is the most common design presently in use on 220-MHz EME.[4,5] Properly duplicated, its gain is about 14.2 dBd. It works best when stacked 8-1/2 feet (2.6 meters) apart in the "E" (horizontal) plane and 8 feet (2.44 meters) apart in the "H" (vertical) plane. If all feedlines are kept short with equal lengths and a good in-phase power divider is used, four of these Yagis will yield close to 22 dBi of gain, the minimum requirement for 220-MHz EME communication as stated above. One commercial manufacturer offers this Yagi with a trigonal reflector.

Several 220-EME stations are using these Yagis stacked 4 wide and 2 high or vice-versa, using the same spacings indicated above. The typical gain is now about 24 dBi; this yields good echoes and improves the performance and ease of working the smaller stations who have met only the minimum EME requirements as specified above.

Some stations are using arrays of the 3.2 wavelength NBS Yagis, but

they are marginal unless at least six are used or the station scheduled has a larger setup to make up the difference for the 1 dB lower gain per Yagi.[4,5] Proper stacking for this antenna is 8 feet (2.44 meters) and 6 1/2 feet (2 meters) in the "E" and "H" planes, respectively.

Another antenna that has been used on 220-MHz EME is the WB6NMT 11-element Yagi.[6] This antenna, a result of the Greenblum designs,[7] is also commercially available. WB0TEM has redesigned this particular antenna by extending the boom about 54 inches (1.37 meters) and adding 3 additional directors for increased gain. Each additional director is spaced 18 inches (45.72 cm) further down the boom and directors 4 through 11 are tapered slightly differently than they were in the original design. WB0TEM and W0SD each use arrays of 16 of these redesigned Yagis at their home stations and also have an 8-antenna array which they have used very successfully on *portable* EME from over 15 states.[8] Recommended stacking for this arrangement is 8 feet (2.44 meters) and 7 1/2 feet (2.29 meters) in the "E" and "H" plane, respectively, and slightly closer (perhaps 6 inches or 15.25 cm) for the shorter version.

A more recent trend is to use the least number of antennas in the array by designing a Yagi with the longest boomlength possible.[9] In this regard, the DL6WU designs are recommended.[10] Indeed, one commercial antenna manufacturer has just announced the availability of a long-boom (30 feet or 9.15 meters) Yagi from these designs.

A few 220-MHz EME stations use parabolic dishes varying from 24 to 42 feet (7.3-12.8 meters) in diameter. Dishes are especially popular with operators who work EME on two or more bands because only the feed system has to be switched to change frequency. Furthermore, if the feed system is properly designed, it can be rotated or switched in polarity to offset the affects of Faraday rotation and thereby improve the chances of a completed EME QSO.

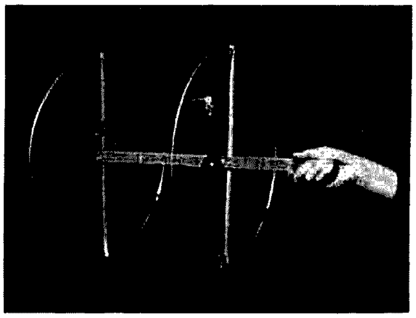

fig. 1. Antenna feed system used by K5FF/W5FF to feed their 30-foot (9.15 meter) dish. (See text for dimensions.)

A 24-foot (7 meter) diameter dish meets the minimum requirement for 22 dBi of antenna gain. As stated in reference 3, an F/D (focal length over diameter ratio) of 0.43 to 0.55 is recommended because it is the easiest to feed. At this frequency, the EIA reference feed is quite large (4 1/2 feet or 1.37 meters square) so other feed systems or variations have been used. WB5LUA modified the EIA reference feed by replacing the screen and driven elements with a pair of 2-element Yagis spaced 1/2 wavelength and placed diagonally on the corners of his 70-cm EIA feed.[3] The reflectors, approximately 5 percent longer than the driven element, are placed 1/4 wavelength behind it. His dish is 24 feet (7 meters) in diameter with a 0.49 F/D ratio. K5FF and W5FF use two circular loops spaced 14 inches (35.6 cm) apart, similar to a quad (**fig. 1**). Each element is made from 1/8-inch (3-cm) diameter wire with the driven element 54 inches (137 cm) and the reflector 56 3/4 inches (144 cm) long. A 1/4 wave 92-ohm Bazooka type balun completes the match to the 170-ohm feed point impedance. Their dish is 30 feet (9.15 meters) in diameter with a 0.43 F/D ratio.

Typical half-power beamwidths for a 22 dBi gain antenna at 220 MHz are 12 to 13 degrees, so some of the medium sized rotators such as the HAM-M series are useable on the smaller antenna arrays. For temporary operation, setting circles and a protractor/level can serve very well for determining true elevation. The later method was used on our recent New Hampshire 220-MHz EME DXpedition (**fig. 2**). However, larger Yagi arrays may require a commensurately stronger and more accurate rotator system such as a prop-pitch motor and selsyns.[3]

receivers

Most 220-MHz EMEers use antenna-mounted very low-noise preamplifiers ahead of a crystal-controlled down-converter. The latter, usually located at the operating position, feeds a suitable high-frequency receiver. The converter should have a noise figure of 2-4 dB maximum with no less than 20 dB image rejection, and should be reasonably free from overload (a potential source of interference is from TV channels 12 and 13).

If you build your own converter, I recommend the use of a high dynamic range preamplifier,[11] a double-balanced mixer,[12] and a 28.1 MHz IF. For 220.0 MHz, use a 191.9-MHz local oscillator derived from a 95.95-MHz crystal oscillator and doubler.[12] There are many commercial converters available. The IF requirements such as stability and bandwidth, among others, are spelled out in detail in reference 3.

Many operators prefer to use transverters on 220 MHz. One of the most popular early arrivals on the 220-MHz scene is the modified Microwave Modules MMT-220/28, available through VE3CRU on special orders. Other transverters are now available from commercial manufacturers. If you prefer to "roll your own" transverter, circuitry is available in references 12 and 13. If you are ambitious, W9SR has recently published a design of a complete transverter for the homebrewer.[14]

preamplifiers

Low-noise GaAs FET (Gallium Arsenide Field-Effect Transistor) preamplifiers are now in wide use on 220 EME operation. Noise figures of less than 1.0 dB are easily obtained even with low cost devices, and this is more than adequate with the sky temperatures prevalent on this band.

One of the most popular circuits is the W6PO design.[15] His original GaAs FET circuits used devices selling for $25.00, but devices such as the Mitsubishi MGF-1202 can easily provide a noise figure of 0.5 dB in this circuit at a cost of less than $10. Other choices are the newer low-cost dual-gate GaAs FETs primarily aimed at the UHF TV market. They will operate in the same circuits providing that an extra intermediate voltage (1.5-3 volts) is applied to the second gate. Some typical low-cost devices are the Mitsubishi MGF 1100, NEC NE41137, Toshiba 3SK121 and the more recent arrival, Motorola MRF966 "sleeper" ($3.55 in a single quantity).[16]

Preamplifier isolation is still a problem at 220 MHz. If you want long life for your input device, a dual relay protection scheme such as the one discussed in reference 3 is highly recommended! Reference 3 also points out that additional spacing is needed between the two cascaded relays if the ultimate isolation is to be obtained.

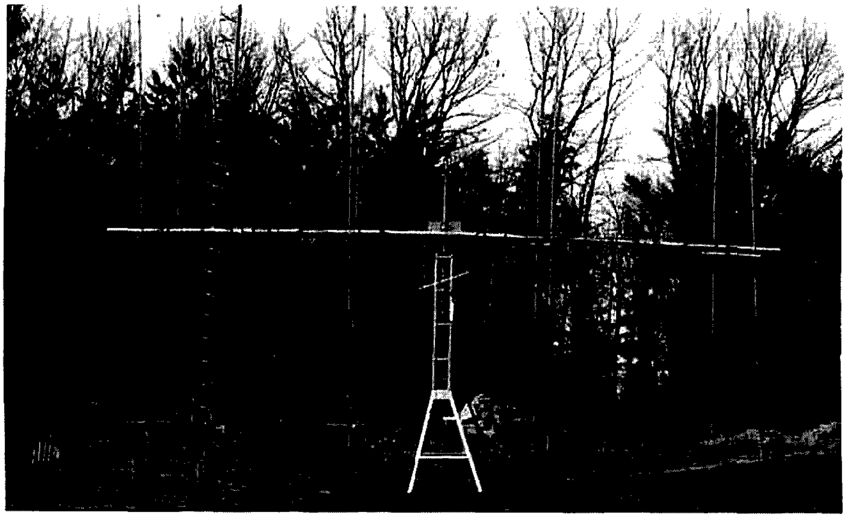

fig. 2. Portable EME antenna system used by W1JR. Eight 4.2λ NBS-type Yagis are mounted on an 11-foot (3.35 meter) high tower. Main boom is a 30-foot (9.15 meter), 3-inch (7.6 cm) diameter irrigation tube.

transmitters

Most 220 EMEers prefer to use an up-converter or transverter driven by a 28 or 50-MHz exciter. This provides the necessary accuracy, flexibility, and stability required for EME. As pointed out in the receiver section of this article, designs are available in references 12, 13, and 14. Commercial transverters such as the Microwave Modules MMT 220/28 and others are available.

The majority of high-power amplifiers used on 220 MHz require greater than 10 watts if full output is required. Most operators either build or buy a moderate power (25-60 watt) solid-state driver. Many circuits have been published and many are available commercially.

Until recently, 4CX250B-type tubes were used in either a modified push-pull plumber's delight with tubing and flapper plates[17] or in the parallel kilowatt unit similar to the K2RIW 70 cm design.[18] 8930 tubes can be substituted for the 4CX250Bs if greater power is required. 8874s[18] or the newly announced EIMAC 3CX800A7 tube should also work well in this design if the input matching network and the output tank circuit are redesigned accordingly, but they will require more drive (40-60 watts) than the 4CX250Bs.

One worthwhile amplifier to consider now available on the surplus market is the AM-6155. Listed as a 50-watt driver manufactured by ITT, it is available from Fair Radio Sales for approximately $150. Included is a coaxial cavity amplifier complete with a tube (8930 typical) and 115 VAC power supply. With a few simple modifications, it will deliver 500-600 watts of output power — not a big EME amplifier, but one to start with and one that's surely more than sufficient for other modes of operation.[19]

The recent (September 1, 1983) FCC change from measuring input to output power and the tantalizing possibility of running legally up to 1500 watts output power has had a profound effect on EME operations. No longer is efficiency a primary requirement. This has led to the use of slightly less efficient but higher output tubes.

In this regard the 8877 is now becoming the most popular tube. It works well using either the W6PO[20] or the ARRL Handbook circuit.[21] (I have been unable to obtain sufficient data from the ARRL to clarify the use of the design in reference 21.) W4WD, who has built it by scaling some of the missing information from the photographs shown, suggests that you extend the Teflon insulation on the plate tank to overlap at least 1/4 inch (6.35 mm) to prevent breakdown. He also experienced some regeneration, but cured it by using the input matching circuitry in the W6PO design. Both of these amplifiers will deliver 1500 watts of output at good efficiency, but require 50 to 60 watts of drive, considerably more than required by typical 4CX250B designs. Finally, most 220-MHz power amplifiers have harmonics that are typically only 25-30 dB below the fundamental. Therefore a harmonic filter such as a series-tuned circuit[18] or a 1/4 wave (at 220 MHz) shorted stub should be placed across the amplifier output connector.

feedlines

Suffice it to say that feedline losses should always be kept to a minimum, especially on EME. A 0.5 dB loss ahead of a 220-MHz preamplifier is acceptable if a very low noise GaAs FET preamp such as the one just described is used. The transmitter feedline loss should be no greater than 1 dB unless you are running the new legal power limit.

Feedline loss is not such a problem

on 220 MHz. Losses are only about 66 percent of those on 70 cm and about 33 percent greater than on 2 meters. Hence, RG-213 coax cable is still usable for phasing lines in a Yagi array. However, 1/2 inch (12.7 mm) or larger hardline is recommended for the transmitter line. Of course, Andrew Corporation Heliax™ is recommended for the ultimate in low loss. The newly announced Belden 9914 is similar to RG-213 with lower loss (approximately 2.5 dB per 100 feet or 30.5 meters at 220 MHz).[22] The Belden 9913 is even lower loss and is similar to RG-213 in size but uses an air dielectric. Either of these coax cables should be useable for phasing lines. A complete list of recommended feedlines is shown in table 1 of reference 3. As stated above, the 200 MHz line loss is approximately 66 percent of the loss listed in the table and the power handling capability is approximately 150 percent of that shown for 70 cm, while the velocity factor remains the same.

system checkout

Assuming that your station is now complete, it's time to go through a checkout to see if all your gear is working properly. The first step is to check VSWR. If it isn't below 1.5:1, and hopefully closer to 1.2:1, it's back to the drawing board. Assuming an acceptable VSWR, the output power should be measured both at the transmitter (to verify FCC requirements) and at the antenna to see that the transmitter feedline loss is low enough. At this point I'll reiterate the necessity of doing power and VSWR tests with an appropriate instrument such as a Bird Electronics model 43 or equivalent. ***Warning: do not stand in front of your antenna with high power applied because your body may absorb hazardous levels of RF radiation.***

Next, check out the receiver. Depending on the noise figure and sky temperature, the audio output from your receiver should decrease from 1 to 2 dB as you elevate your EME antenna from the horizon to directly overhead. You may also move the overhead point several degrees one way or the other to find a "cold" spot which will yield the maximum drop in receiver output noise. Next, aim your antenna at the sun after noting the receiver output on the cold sky. The noise level should increase from 7-10 dB depending on your antenna size, noise figure, and the conditions on the sun at the time of measurement. Check with another EMEer to compare figures that are presently being measured. (This procedure is described in more detail in reference 3.)

Now you can check for echoes; don't get frustrated if you don't hear any. (The Faraday rotation may not be cooperating.) Remember that the round-trip path to the moon is just over 2.5 seconds, so long dahs or letters may be sent for test purposes. Also, doppler shift can be up to ± 750 Hz on 220 MHz. The maximum will be up to $+750$ Hz at moonrise and -750 Hz near moonset, with little or no doppler when the moon is directly south of your QTH. If Faraday is uncooperative, wait 30 minutes to an hour and try again. Better yet, get another EMEer to either listen for you or transmit a test signal for you to listen for.

scheduling

220 MHz EME operation is usually conducted between 220.005 and 220.050 MHz with 220.020 MHz as the calling or CQ frequency. DXpeditions usually operate on 220.035 MHz and listen 5-10 kHz up. The activity is usually centered near perigee similar to the 70 cm EME weekend. To allow for those who work multiple bands, 220 MHz EME is usually conducted in the Saturday evening/Sunday morning time frame.

About a year ago, the 220 MHz EMEers decided to adopt the 70 cm scheduling and reporting system. Hence, the transmission/receiving periods are 2 1/2 minutes long, with the westernmost station (and DXpeditions) transmitting the first 2 1/2 minutes of each 5-minute block. A "T" report indicates signals or letters received. An "M" report means that positive identification (both call signs) has been received, and is therefore valid for a contact, while an "O" report signifies "Q5" copy. The usual acknowledgements both ways with "R" completes the exchange.

No formal 220 MHz scheduling is conducted as on 2 meters or 70 cm, so you will have to set up your own schedules until activity warrants more formal arrangements. However, the 70-cm EME net is often used to set up 220-MHz EME schedules. You can find this net between 1600-1700 UTC every Saturday and Sunday on 14.345 MHz. Many EMEers also meet to exchange schedules and information by using OSCAR 10 on a downlink frequency of 145.950 MHz when the satellite is in view.

summary

You too can join the fun on a frequency that in many ways is less critical or demanding than 2 meters or 70 cm yet still challenging. I have tried to sum up the state-of-the-art on 220 MHz and provide a recipe for simple EME success. Because 220-225 MHz antenna systems are often smaller than those used on 2 meters, the lower sky temperature and the availability of suitable designs or commercial equipment make 220 MHz an ideal band for conducting EME schedules and experiments. There is also the challenge of attaining WAS on a band where this feat has only been recently accomplished. Good luck — see you on 220 soon.

acknowledgements

As in past columns, I have had to rely on many persons to provide some of the material needed to put this column together. I'd particularly like to thank Lewis Collins, W1GXT; Fred Merry, W2GN; Ron Barlow, N4GJV; Russ Wicker, W4WD; Lee Fish, K5FF; Fred Fish, W5FF; Al Ward, WB5LUA; Ed Gray, WØSD; and Marc Thorson, WBØTEM for all the help and encouragement they gave me to make this month's column possible.

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VHF/UHF coming events

September 8,9: *ARRL VHF QSO Party*
September 25: *EME perigee*

DX news note

The 23 cm (1296 MHz) tropo record was broken on June 24, 1984, at 0035 UTC. N6CA at 1100 feet (335 meters) on the Palos Verdes Peninsula in Torrance, California, running 150 watts output and one 44-element loop Yagi, worked KH6HME at 8200 feet (2500 meters) ASL on the eastern slope of Mauna Loa, Hawaii, running 20 watts output and four 25-element loop Yagis. The opening, while lasting several hours on 2 meters (where liason was conducted), lasted only about 10 minutes on 23 cm. This extends the DX record on this frequency to 2472 miles (3977 km).

ham radio

improving amplifier ALC circuits: part 2

MLA-2500 input matching and tube protection circuits

The **MLA-2500** has acquired a reputation as a tube eater, and it's no secret that the cost of replacing the final tubes in the MLA-2500 has risen from $75 to $520 in just ten years. This has left many MLA-2500 owners nervous about — if not fearful of — tube loss. The ALC and driver matching circuit described in this article should significantly decrease the possibility of tube loss due to overdrive.

One problem that has plagued many MLA-2500 owners is the lack of an input matching circuit. A Kenwood TS-430, for example, is almost unusable with the MLA-2500. The success of the E.T.O. "Alpha" amplifier series has allowed several design ideas to be field-proven. I adapted the broadband, untuned input circuitry used in the small E.T.O. amplifiers to cure the input mismatch problem. The matching circuit is simple to design and construct. Ten trifilar turns of No. 18 enameled wire are wound on a T94-2 core. An electric drill is used to twist three 15-inch lengths of wire together. (The completed coil is shown in **figs. 1** and **2**. **Fig. 3** shows the original input mounting around the input relay in an MLA-2500B. **Fig. 4** shows the proper mounting of the input toroid. The wirewound "Non-inductive" swamping resistor used in the MLA-2500 was found to be too reactive to be of use and was removed. I constructed a 50-ohm, 40-watt non-inductive resistor suitable for swamping use from twenty 1-kilohm, 2-watt carbon resistors. I tightly packed the resistors into two groups of 7 and 13 respectively (shown in **figs. 5** and **6**) and then mounted them as shown in **fig. 7**. The interconnection is shown schematically in **fig. 8**.

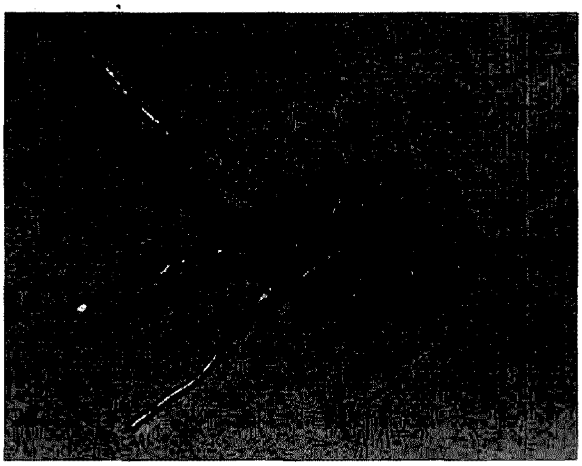

fig. 1. Toroid input transformer consists of 10 trifilar turns on a T94-2 core.

The next step in stabilizing the input impedance involved moving the cathode RF choke, RFC-4, from the ALC board, PC-1004, to the cathode area. Some older MLAs use an unbypassed RF choke, RFC-7, already installed in series with the cathode line to the ALC board. If no RFC-7 exists, remove RFC-4 from the ALC board and reconnect it between the cathode line common and an empty terminal on the barrier strip mounted between the tube sockets. Install a jumper between the ALC board terminals from which RFC-4 was removed. Move the cathode return line to the barrier strip terminal and add a 0.01 μF bypass capacitor to ground as shown in **fig. 8**. After installation, the MLA-2500 will present less than a 1.3:1 VSWR to the exciter on all bands from 80 to 10 meters; on 160 meters the VSWR is 1.3:1. Although some variation in the value of the compensating capacitor, C_C, may be required, 100 pF connected as shown in **fig. 2** has proven optimum in all retrofits completed to date. The use of different core material and/or core size is likely to necessitate some variation in the value finally selected and the connection terminal.

By J. Fred Riley, WA8AJN, 1721 Poplar Street, Kenova, West Virginia 25530

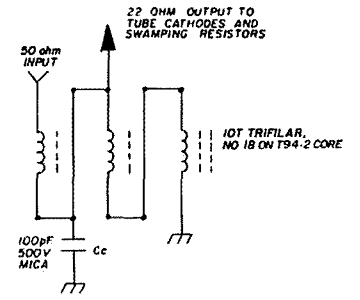

fig. 2. Toroid internal and external connections show position of compensating capacitor.

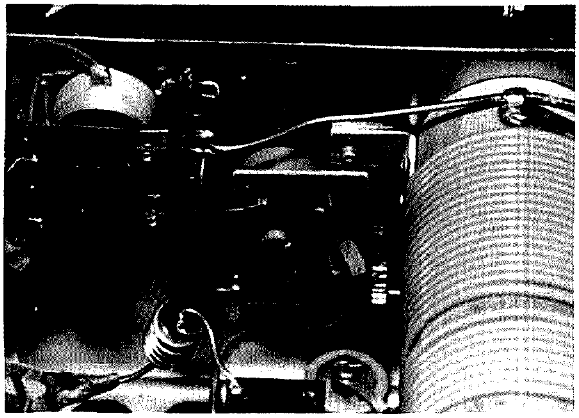

fig. 3. MLA-2500B input relay area before modification.

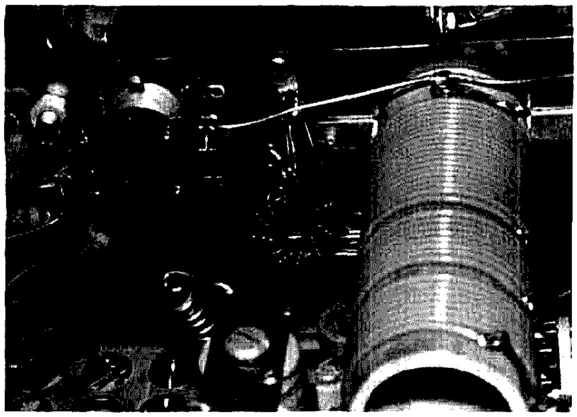

fig. 4. MLA-2500B input relay area shows toroid connections and mounting to ground lug and relay terminal. Compensating capacitor is visible at top of toroid.

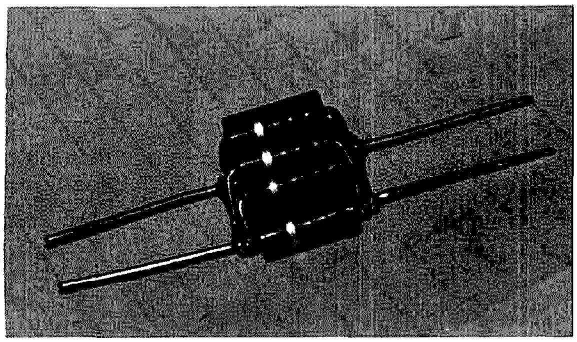

fig. 5. Construction technique for seven-resistor pack.

ALC

The final modification to the MLA-2500 may well be the most important and beneficial to improved performance. The technique employed is a modified version of the circuit used in the E.T.O. Alpha 77. It is a grid-current-derived ALC circuit that uses readily available Radio Shack parts. I could not use the original Alpha 77 circuit because it uses a positive sample voltage whereas negative is used in the MLA-2500. In installing the ALC circuit I also remounted the thermal sensor and increased the voltage to the cooling fan.

I remounted the thermal sensor, SW-3, shown in **fig. 9**, from behind the 40-meter loading padder, C42, to behind the 80-meter loading padder, C46. The final position is shown in **fig. 10**. Why the inside tube should run hotter is not obvious. Nevertheless, the inside tube *does* run hotter, and moving the thermal switch makes sense. Perhaps the bandswitch disturbs the airflow. You may wish to replace the cooling fan; after extensive evaluation of cooling fans, the Rotron WR2A1 "Whisper" is highly recommended. It is both quiet and efficient. I also removed R20, a 500-ohm, 10-watt resistor, from the thermal sensor circuit. When I removed R20 and rewired around it, I saved the thermal strip and remounted it to the bottom of the chassis below the 40-meter loading capacitor, C42, using C42's mounting screw. This terminal strip is used in the ALC circuit and is shown at the top center of **fig. 7**. Fifty milliamperes, the new maximum grid current, represents only one-twentieth scale on the original MLA-2500 meter. By replacing the grid shunt, R19, with a 10-ohm, 2-watt resistor and adding a 1-kilohm, 1/4-watt resistor in series with the grid meter line, the full-scale meter reading is increased to 100 mA. As a benefit, enough voltage is available across the new grid shunt to activate the new ALC circuit shown in **fig. 5** of part 1. (**Fig. 6** of Part One of this article shows the construction technique used and **fig. 7** of Part One shows the mounting to the pushbutton mounting nuts.)

I used the ground lug on the remounted terminal strip to provide a ground for the new grid shunt, R19, and to provide a terminal point at which to connect

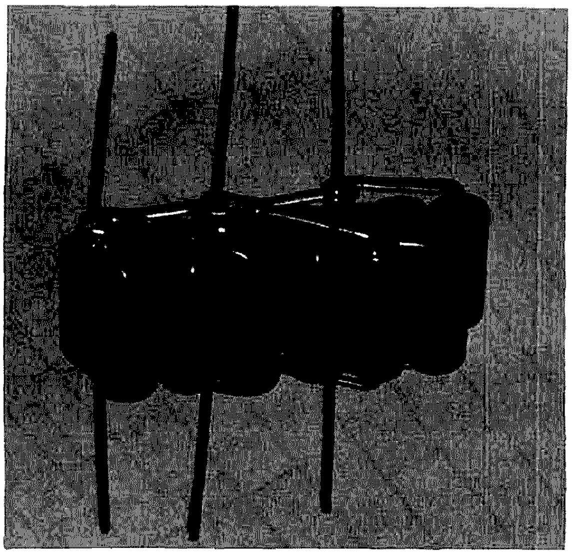

fig. 6. Construction technique used for thirteen-resistor pack.

the grid meter line to the new series metering resistor, a 1-kilohm, 1/4-watt resistor. A new wire was added inside the wiring harness from the grid line to the area of the function switch mounting nuts. Using terminal strips modified in Dentron fashion and discrete components I constructed the basic circuit shown before. The −18 VDC required was obtained by a simple modification to the power supply board, PC-1002. Later model MLA-2500s do not have the 120 VAC winding shown on the schematic. If your unit has a black wire connected to pin No. 3 on the bottom of power supply board, PC-1002, unsolder the wire and tape it back. If your unit has a C32 capacitor installed, remove it. Remove the line going to pin No. 1 bottom — if it exists — and pull it back to the ALC circuit construction area. If no line exists between the ALC and power supply boards, run an additional wire in the harness to connect the ALC board to the new ALC circuit.

On the power supply board, PE-1002, install the new C32 capacitor, observing polarity. Diode CR16 may or may not be present. Use the old CR16 or any general purpose power diode. It may be most convenient to tack solder CR16 in from the 1/T, 2/B (12 VAC) trace on the power supply board to the 1/B terminal trace to which C32 is soldered, refer to part 1, Fig. 5 for schematic representation. Carefully locate the lines to the transmit light, X-2, and unsolder them from the power supply board. You may wish to use part of the excess wire connecting X-2 to the new ALC circuit for the new line connecting the 1/B (−18 VDC) terminal to the ALC circuit. After connecting X-2, the power supply board, and the new ALC circuit, only the ALC board, PC-1004, remains to be modified.

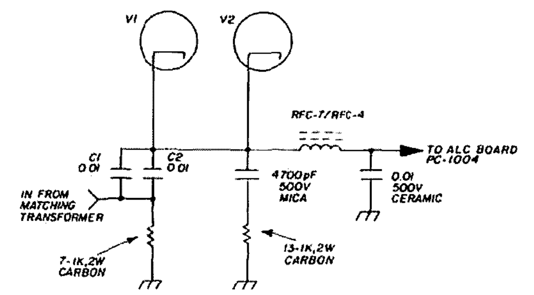

fig. 7. Schematic representation of connections from tube cathodes to new swamping resistor packs and remounted cathode choke.

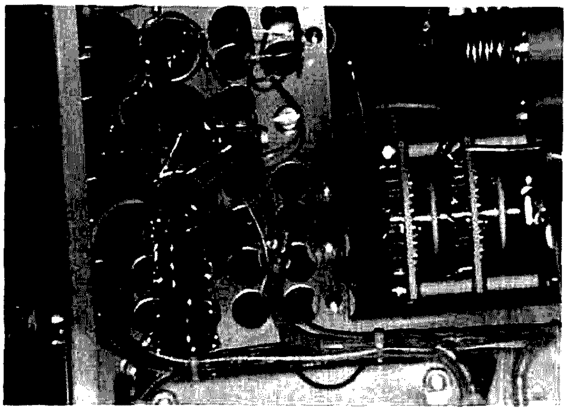

fig. 8. Installation of resistor packs clearly shown in spaces between tube sockets and wiring harness. Note remounted terminal strip for grid shunt and meter multiplier at top right of photograph.

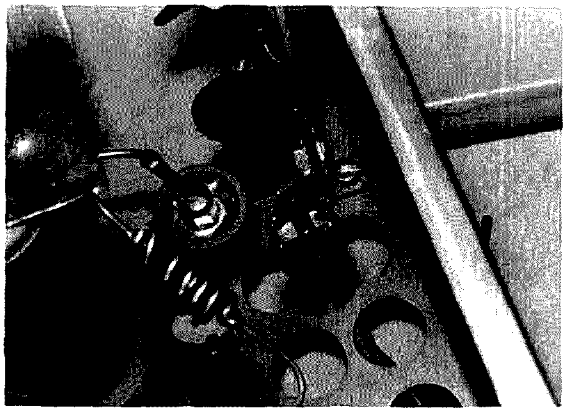

fig. 9. Original mounting of thermal sensor.

modifying the ALC board

Remove C22, RFC3 (if present), and R10 from the ALC board, PC-1004. Replace R12 with a 1-kilohm,

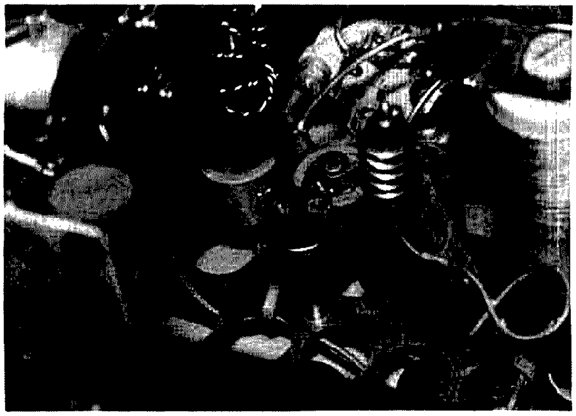

fig. 10. Remounted thermal sensor positioning.

1/4-watt resistor. Tack solder the old R12, a 27-kilohm, 1/4-watt resistor, from the input, pin No. 2, trace to ground. Using R10's terminals, correctly remount CR18 in the circuit. This completes the modifications. Check the ALC circuit by insuring that the transmit light, X-2, comes on at 55-60 mA of grid current; −8 VDC should appear at the ALC jack simultaneously with X-2's illumination. The new ALC circuit provides protection from accidental or transient overdrive when connected to the exciter. The input swamping and matching circuit helps limit the maximum grid current available. The increased sensitivity of the grid meter and X-2's visual indication prove powerful tools in preventing accidental or transient overdrive. The ALC circuit should act in concert with the change in cooling to protect the tubes under almost any operating conditions.

"Inadequate warm-up time is something these modifications cannot protect against. Four or five minutes has been found to be a practical minimum. Internal tube arcing can occur even though Eimac's specification sheet calls for sixty second minimum and the MLA provides seventy-five seconds. Increasing the time for warm-up can unquestionably save your tubes."

acknowledgment

I wish to thank Tom Keadle, W8EII, for the photographs and Rodger Miller, KC8DA, for the use of his amplifier.

ham radio

ham radio TECHNIQUES

Bill Orr W6SAI

shortwave circus

As time goes by, more and more modern HF transceivers come off the production line with the capability to receive the range of 100 kHz to 30 MHz built right in. This has opened up a whole new world to Amateurs who thought that the shortwave spectrum ended at the edges of the ham bands.

It's fun to tune the spectrum; no wonder the SWL hobby has grown so rapidly over the past few years. There are lots of interesting signals to hear in the HF range!

The most obvious signals come from the shortwave broadcast stations; there are plenty of them, and most of them are listed by frequency and operating hours in the *World Radio Handbook*.[1] Even the casual observer will find other signals: point-to-point transmissions, ship-to-shore, aircraft, military, RTTY, FAX, and so on. SWL magazines provide, on occasion, lists of these interesting transmissions.

In a few days' listening time, you can gain a good idea of what is going on outside the ham bands, as far as legal, recognized transmission goes. But an even more interesting field to explore is that category of transmissions that are undercover, clandestine, or modified in such a way that the casual listener cannot comprehend the transmitted information. And then, there are other signals . . .

things that go bump in the night

The shortwave spectrum is full of incomprehensible and confusing signals that whet the interest of the serious listener. Some of these signals are quite loud and easily found.

Sweepers. One of the most common "mystery" signals is the sweeper. This is an unsteady carrier, loud and rough. It "sweeps" through your receiver quite rapidly, as though someone had tuned a wobbly VFO across your listening frequency. These signals, most noticeable in the 14 and 26 MHz regions of the spectrum, are in fact emitted from large, industrial heat-treating machines used in the plastics industry, making water beds, furniture, dishes, and other sundries. These industrial RF oscillators run upwards of 100 kW and shift frequency as the manufacturing process advances. While they *should* be in the ISM (Industrial-Scientific and Medical) assignments at 13.56 MHz and 27 MHz, they seem to roam the spectrum at will.

I spoke to the operator of such an oscillator once. He told me that he was an avid CBer who made sure his oscillator never landed in the CB channels (that's a Goody Buddy, for you!).

Many other industrial RF oscillators exist for various purposes, and most of them radiate – sometimes badly. They're often operated by untrained personnel who have little concern for radiation. The minimal shielding on these machines is frequently removed to facilitate easy loading and unloading of the material they are designed to process.

Woodpeckers. Most hams are familiar with the obnoxious woodpecker signals that infest the Amateur bands. They exist in force too, on other frequencies. At least four such signals exist; they are over-the-horizon, high power backscatter radars operated by the Soviet Union. With the United States as the target, the radars search for telltale missile trails and other interesting reflections of military importance to the Soviet radar operators.

Repeated protests about the disruptive effects of these long-distance radars have been lodged with the USSR. The extreme power of these devices, plus the sharp wavefront, make them a pest to all users of the HF spectrum.

Five-letter "numbers" groups. It won't take you long to find the 5-letter mystery stations. These can be SSB, AM, or CW signals that transmit code groups at length, with no recognizable identification. On voice, the transmissions may be in English, Spanish, or German. The voice repeats 5-letter code groups of numbers or letters. The most numerous are the "numbers" stations. A typical message sounds like: "83457-90030-45089-10019-63345," and so on; often the message is repeated again and again. No station identification is ever given. Some stations repeat a series of letters in the phonetic alphabet instead of numbers, and others mix numbers and letters. Who are these stations? Where are they? And what are they doing?

"Spook" transmission. Harder to find are the two-way "spook" signals. (A good place to look for these is just above the 10 MHz Amateur assignment.) You'll hear two-way CW conversations at about 8 to 10 WPM (very

poorly sent, by the way). A typical transmission will run: "AJ23 AJ23 DE 64Z QRK?" Then the other station will reply, "64Z DE AJ23 QRK 5 GA." Then the coded message will start. (Just after the 10 MHz assignment opened for Amateurs, I ran across a pair of spooks trying to work each other. They were both S7 but they had a terrible time, stumbling around, trying to make contact. I couldn't resist, and broke in with a "QRK 5 GA PSE." This *really* upset the spooks. They frantically signalled each other, and in a burst of poorly sent CW, moved out of the Amateur assignment.) Careful listening will reveal a myriad of these signals, many of them coming from Central America.

Scientific sounders. You'll find a lot of more routine signals, too. Ionospheric sounders run up and down the spectrum, sounding like a string of dots passing quickly across your listening frequency. Others are more sophisticated and have a more complex sound. An interesting trio of stations used to monitor potential earthquake activity in California can be found on 5.115 MHz, 3.395 MHz, and 10.163 MHz. Located along a major fault, each station transmits a steady carrier (without identification), and the signals are observed at a monitoring point in Utah. When an earthquake occurs, the ionosphere above the quake area is disturbed; by closely observing the signals, investigators can learn much about the relationship between earthquakes and the accompanying ionospheric disturbance.

In this same vein, certain obscure data is transmitted by some AM broadcast stations that employ a 20 Hz FSK on their carrier. One station that does this is KNX in Los Angeles (1070 kHz).

The single-letter beacons (SLB). Of great interest to some listeners are the so-called single-letter beacons, which have been on the air for up to 20 years with little publicity. One or two of these are in the ham bands.[2] On the west coast, in the early morning hours when conditions are good, the SLB on about 3979 kHz is quite loud. It simply sends the Morse letter K about every four seconds in FSK. It can be quite annoying to early-morning nets on this channel.

Many other SLBs exist across the HF spectrum. Some are on-off keyed, others are FSK. Many of the latter use the Eastern European shift of 1 kHz, which suggests Soviet origin. The SLBs send different individual letters, and sometimes will send a short burst of five number code groups at about 20 WPM, then return to the carrier or the keyed letter. A large number of beacons, fading in and out with the skip, have been logged in the United States.

One set of beacons is particularly interesting, since they are well received on the west coast. This set comprises a number of *K* beacons. The beacons are keyed FSK simultaneously and are on 9.043, 11.156, 12.152, 14.478, 14.968, 18.086, and 18.349 kHz (approximately). Signal arrival indicates that the beacon set is located in Siberia, possibly on the Kamchatka penisula.

There are plenty of SLBs on the air, as the footnote indicates, and they seem to be heard all over the world. What is their purpose? Where are they? What messages do they convey? And who are the recipients? Your guess is as good as mine.

"Cut-number" stations. Some unidentified CW stations, in addition to sending coded messages, encode the numbers in the message. The transmissions sound like letter groups when they are really number groups. For example, the digit 1 is sent as A. 2 as U, 3 as V, 5 as E, and so on. This can be quite confusing to the casual listener.

Spread-spectrum signals. Various forms of wideband transmissions can be logged in the HF spectrum. One subtle form sounds like "white noise" on a receiver. This hiss occupies about 100 to 200 kHz of spectrum space. For some time such a signal was apparent in the high frequency end of the 20-meter phone band, but nobody seemed to notice it. Along with the Swedish Amateur who first pointed it out to me, I ran a directional plot on it and found that it seemed to be coming from England, as far as we could determine. In recent years, the signal has moved out of the 20-meter band.

Another form of spread-spectrum transmission is noticed by the perceptive observer as bits and pieces of voice transmission that occasionally pop up on your receiver. Each burst is very short; only fractions of words can be heard. The signal is obviously jumping around at a very fast rate!

"Junk" signals. A few minutes listening to the HF spectrum reveals an amazing quantity of sloppy signals. It seems as if all the old World War II surplus transmitters must be on the air somewhere. For years Amateurs on the west coast were plagued by slurpy, burpy coded CW signals that jammed the 80-meter band during the early morning hours. Most of the transmissions were in Chinese. Some are still on the air. Many South and Central Americans use ham gear as telephone links between isolated locations. The 10-meter FM channels are occasionally blocked by Spanish-language signals carrying on telephone-type conversations.

A quiet scandal (seemingly ignored by the FCC) is the proliferation of illegal CB-type operations between 27 and 28 MHz. The casual listener will soon pick out loud signals, some of whom run kilowatts of SSB power into large beam antennas. Not long ago one such illegal operator boasted that he had worked over 90 countries with his illegal transmissions.

Another collection of "junk signals" exists just outside the low frequency end of the 160-meter band. These are the wireless telephones which transmit frequency modulated signals in the span of 1650 kHz to 1800 kHz. Some of them operate right up into the low end of the 160-meter Amateur assignment. With a good antenna in a quiet location, a wireless phone can be heard for up to 10 miles. I wonder if the users of these phones know that they are

furnishing amusement to casual eavesdroppers?

"Free-radio" broadcasting. Do you like to play music on the air? Join the free-radio broadcasters and play hide-and-seek in the radio spectrum! One popular frequency for pirate broadcasters is 1605 kHz, just outside the top end of the broadcast band. Others pop up on various frequencies near the short wave broadcast bands, and a few operate in the 88-108 MHz FM band. Although pirate broadcasters are quite rare in the United States, many exist in Europe. Because they vary frequency and time of broadcast, they are difficult to pinpoint, but many send QSLs for reception reports! The FCC is quick to crack down on pirate broadcasters, and it is not easy to spot one operating in the United States.

Soviet jammers. One of the major occupants of the international broadcasting bands transmits only noise, in a deliberate attempt to prevent listeners from hearing the program material on certain frequencies. It has been estimated that there are more than 2,000 jamming transmitters, located principally in the USSR, Bulgaria, Czechoslovakia, and Poland, aimed at the local language broadcasts beamed behind the "Iron Curtain" from the west. Illegal or not, they are a reality. Some sound like a buzzsaw, others a rhythmic whine. Many are quite broad. The majority have identification signals, such as ZG or U7 which are sent in slow-speed Morse.

To make matters more interesting, Soviet broadcasts to China are jammed by the Chinese. Sometimes this jamming takes the form of taped music played backward! What next?

Illicit drug trade. Because HF radio is widely used by drug smugglers, the alert listener can occasionally run across transmissions dealing with this underground activity. These messages, mainly on SSB, are generally in English or Spanish. Nicknames are used instead of calls, and most of the traffic seems to occur between 4 MHz and 14.5 MHz, often just outside an Amateur band.

Good stuff, too! Aside from this long list of assorted follies and undesirables, there are many other interesting things to monitor outside the Amateur bands: point-to-point RTTY for the news services of the world, airline networks, weather broadcast, military, MARS (Military Affiliate Radio Stations), Coast Guard, and INTERPOL. The list goes on and on. The point is that there's plenty going on outside the Amateur bands.[3] Exploring these regions is a fascinating undertaking. You'll be suprised at what you can hear! Good listening!

references

1. J.M. Frost, Editor, *World Radio and TV Handbook*, 38th Edition, available from Ham Radio's Bookstore, Greenville, N. H. 03048 ($20 postpaid).
2. Observed frequencies of some of the SLBs heard on the west coast are (in MHz): 3.979, 4.006, 6.227, 7.512, 7.557, 8.138, 8.146, 8.646, 9.043, 9.058, 10.645, 11.156, 12.152, 12.185, 12.329, 13.329, 13.638, 14.478, 14.968, 17.016, 18.016, and 18.349 (all frequencies plus or minus 2 kHz).
3. For more information on activity in the radio spectrum, see *Popular Communications*, published monthly by Popular Communications Publishing Group, 76 North Broadway, Hicksville, New York 11801.

ham radio

packet radio: the software approach

AX.25, 1200 baud packet on the TRS-80 Models I and III

There have been a number of excellent articles[1,2] published recently on the new Tucson Amateur Packet Radio (TAPR) AX.25 protocol terminal node controller (TNC). This TNC communicates with a host microcomputer over an RS-232C or parallel interface and includes its own 6809 microprocessor, 32K of EPROM, 8K dynamic RAM, an SDLC/HDLC controller, an RS-232 UART, and a number of ancillary support chips. It is, in essence, a complete dedicated packet microcomputer.*

There is yet another approach to 1200 baud synchronous packet using the AX.25 protocol: a software approach for the TRS-80 Model I or Model III that eliminates the TNC's dedicated 6809 microprocessor, SDLC/HDLC controller, 32K of EPROM, 8K of RAM, RS-232C UART, and all ancillary support chips.

Like the TAPR TNC, the software approach requires a 1200 baud modem using the EXAR 2206/2211 AFSK modulator/demodulator chips. It also requires a port zero encoder/decoder as an interface between the TRS-80 microcomputer and the outside world. Both the modem and port zero encoder/decoder may be homebrewed for a total cost for parts of approximately $25 to $30.

The balance of this article describes the three fundamental software subroutines used to do the following:

*Available in kit form from the TAPR group, $247 prepaid. The kit is of superb quality and workmanship and may be assembled by the average Radio Amateur with modest kit building experience in eight to ten hours. Contact Tucson Amateur Packet Radio, Inc., P.O. Box 22888, Tucson, Arizona 85734 for information.

- convert the computer's parallel 8-bit byte to a 1200 baud synchronous serial data stream with real-time zero insertion where necessary during transmission.
- convert the 1200 baud incoming synchronous data stream to parallel 8-bit bytes with zero deletion in real-time during reception.
- virtual real-time cyclic redundancy checking (CRC) of incoming packet frames and CRC generation for transmitted packet frames.

1200 baud transmit parallel byte-to-serial bit conversion

This is illustrated in the relatively simple subroutine in fig. 1's commented source code. All opening and closing flags are generated by **CALL**ing **SN1A** with the unique 126 decimal flag byte in the 'A' register. Call SN1A at least 60 or 70 times for the first frame's opening flags, once for frame separator flags (in a multi-frame packet the closing flag for one frame serves as the opening flag for the next frame), and call it once for the entire packet's closing flag.

To transmit data between opening and closing flags that may require zero insertion, CALL SN1 with the data byte in the 'A' register. Either SN1A or SN1 must be called sequentially with a maximum delay between calls of about 40 microseconds so as not to disrupt the timing of the serial synchronous data stream.

1200 baud receive serial bit-to-parallel byte conversion

Figure 2's commented source code shows the subroutine that converts the incoming 1200 baud synchronous bit stream to parallel 8-bit data bytes with zero deletion where necessary. Most of the Z-80's regular and alternate registers are used by the IN1 routine (originally created by W2EUP). This routine is

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fig. 1. 1200 baud SDLC/HDLC fundamental transmit subroutine.

```
00100 ;                          FIGURE 1
00110
00120 ; 1200 BAUD SDLC/HDLC FUNDAMENTAL TRANSMIT SUBROUTINE
00130
00140 ; FLAGS ARE TRANSMITTED BY CALLING SN1A WITH 126 IN
00150 ; THE 'A' REGISTER......NO ZERO INSERTION IS REQUIRED.
00160
00170 ; DATA BYTES THAT MAY REQUIRE ZERO INSERTION ARE TRANS-
00180 ; MITTED BY CALLING SN1 WITH THE DATA IN 'A' REGISTER.
00190
00200 ; LINE 1620 'SPEED' SHOULD = 98 DECIMAL FOR THE MODEL
00210 ; I AND 115 DECIMAL FOR THE MODEL III'S FASTER CLOCK.
00220
00230
00240 SN1     LD      D,A             ;BYTE VALUE TO TRANSMIT
00250         LD      E,8             ;NUMBER OF BITS PER BYTE
00260 SN2     LD      A,(LASONE)      ;1 = SPACE & 5 = MARK
00270         CP      1               ;WAS IT A SPACE ?
00280         JP      Z,LASTSP        ;IF SO, GOTO LAST SPACE
00290         BIT     0,D             ;SET Z FLAG FOR BIT ZERO
00300         CALL    NZ,MARK         ;IF NOT ZERO SEND MARK
00310         BIT     0,D             ;SET Z FLAG FOR BIT ZERO
00320         CALL    Z,SPACE         ;IF ZERO SEND SPACE
00330         NOP                     ;2 USEC TIMING ADJUST
00340         DEC     E               ;-1 FROM BIT COUNTER
00350         RET     Z               ;IF ZERO, GET NEXT BYTE
00360         RRC     D               ;RIGHT SHIFT ALL 1 BIT
00370         JP      SN2             ;GO BACK FOR NEXT BIT
00380 LASTSP  BIT     0,D             ;SET Z FLAG FOR BIT ZERO
00390         CALL    NZ,SPACE        ;IF NOT ZERO SEND SPACE
00400         BIT     0,D             ;SET Z FLAG FOR BIT ZERO
00410         CALL    Z,MARK          ;IF ZERO SEND MARK
00420         NOP                     ;2 USEC TIMING ADJUST
00430         DEC     E               ;-1 FROM BIT COUNTER
00440         RET     Z               ;IF ZERO, GET NEXT BYTE
00450         RRC     D               ;RIGHT SHIFT ALL 1 BIT
00460         JP      SN2             ;GO BACK FOR NEXT BIT
00470 SPACE   LD      A,5             ;SEND SPACE TONE
00480         OUT     (0),A           ;VIA PORT ZERO
00490         XOR     A               ;ZERO OUT 'A' REGISTER
00500         LD      (ZEROMK),A      ;AND ZERO NARK COUNTER
00510         LD      A,(SPEED)       ;COUNTDOWN VALUE
00520         LD      HL,SPACEA       ;RETURN MEM LOCATION
00530         PUSH    HL              ;PUSH ON TOP OF STACK
00540         LD      HL,DECSP        ;JP (HL) ADDRESS
00550 DECSP   DEC     A               ;-1 COUNTDOWN VALUE
00560         RET     Z               ;GOTO SPACEA WHEN ZERO
00570         JP      (HL)            ;JUMP TO DECSP
00580 SPACEA  LD      A,(LASONE)      ;PREVIOUS BIT SENT
00590         CP      5               ;WAS IT A MARK ?
00600         JP      Z,SPACEB        ;IF SO, DON'T COUNT IT
00610         LD      A,(ZEROSP)      ;SPACE COUNTER STASH
00620         INC     A               ;+1 TO SPACE COUNTER
00630         CP      5               ;5 SPACES IN A ROW ?
00640         JP      Z,SPACFC        ;IF SO, DO ZERO INSERTION
00650         LD      (ZEROSP),A      ;IF NOT, SAVE NEW VALUE
00660         NOP                     ;2 USEC TIMING ADJUST
00670         RET                     ;RETURN WHENCE U CAME +1
00680 SPACEB  LD      A,1             ;SINCE NOT SAME CHANGE IT
00690         LD      (LASONE),A      ;UPDATE LASTONE
00700         NOP                     ;EQUALIZING DELAY
00710         NOP                     ;EQUALIZING DELAY
00720         NOP                     ;EQUALIZING DELAY
00730         RET                     ;RETURN WHENCE U CAME +1
00740 SPACEC  LD      A,1             ;1 = SPACE & 5 = MARK
00750         LD      (LASONE),A      ;UPDATE LASTONE
00760         LD      BC,1            ;DELAY - NO SN2 ITERATION
00770         CALL    060H            ;APPROX. 30 MICROSECONDS
00780         CALL    MARK            ;DO ZERO INSERTION
00790         XOR     A               ;ZERO OUT 'A' REGISTER
00800         LD      (ZEROMK),A      ;AND ZERO MARK COUNTER
00810         RET                     ;RETURN WHENCE U CAME +1
00820 SPACE1  LD      A,5             ;SPACE1 ONLY FOR FLAG
00830         OUT     (0),A           ;SPACE TONE PORT ZERO
00840         LD      A,1             ;1 = SPACE & 5 = MARK
00850         LD      (LASONE),A      ;UPDATE LASTONE
00860         XOR     A               ;ZERO OUT 'A' REGISTER
00870         LD      (ZEROMK),A      ;AND ZERO MARK COUNTER
00880         LD      A,(SPEED)       ;COUNTDOWN VALUE
00890         LD      HL,DECSP1       ;JP (HL) ADDRESS
00900 DECSP1  DEC     A               ;-1 COUNTDOWN VALUE
00910         RET     Z               ;RETURN WHENCE U CAME +1
00920         JP      (HL)            ;JUMP TO DECSP1
00930 MARK    LD      A,1             ;SEND MARK TONE
00940         OUT     (0),A           ;VIA PORT ZERO
00950         XOR     A               ;ZERO OUT 'A' REGISTER
00960         LD      (ZEROSP),A      ;AND ZERO SPACE COUNTER
00970         LD      A,(SPEED)       ;COUNTDOWN VALUE
00980         LD      HL,MARKA        ;RETURN MEM LOCATION
00990         PUSH    HL              ;PUSH ON TOP OF STACK
01000         LD      HL,DECMK        ;JP (HL) ADDRESS
01010 DECMK   DEC     A               ;-1 COUNTDOWN VALUE
01020         RET     Z               ;GOTO MARKA WHEN ZERO
01030         JP      (HL)            ;JUMP TO DECMK
01040 MARKA   LD      A,(LASONE)      ;PREVIOUS BIT SENT
01050         CP      1               ;WAS IT A SPACE ?
01060         JP      Z,MARKB         ;IF SO, DON'T COUNT IT
01070         LD      A,(ZEROMK)      ;MARK COUNTER STASH
01080         INC     A               ;+1 TO MARK COUNTER
01090         CP      5               ;5 MARKS IN A ROW ?
01100         JP      Z,MARKC         ;IF SO, DO ZERO INSERTION
01110         LD      (ZEROMK),A      ;IF NOT, SAVE NEW VALUE
01120         NOP                     ;2 USEC TIMING ADJUST
01130         RET                     ;RETURN WHENCE U CAME +1
01140 MARKB   LD      A,5             ;SINCE NOT SAME CHANGE IT
01150         LD      (LASONE),A      ;UPDATE LASTONE
01160         NOP                     ;EQUALIZING DELAY
01170         NOP                     ;EQUALIZING DELAY
01180         NOP                     ;EQUALIZING DELAY
01190         RET                     ;RETURN WHENCE U CAME +1
01200 MARKC   LD      A,5             ;1 = SPACE & 5 = MARK
01210         LD      (LASONE),A      ;UPDATE LASTONE
01220         LD      BC,1            ;DELAY - NO SN2 ITERATION
01230         CALL    060H            ;APPROX. 30 MICROSECONDS
01240         CALL    SPACE           ;DO ZERO INSERTION
01250         XOR     A               ;ZERO OUT 'A' REGISTER
01260         LD      (ZEROSP),A      ;AND ZERO SPACE COUNTER
01270         RET                     ;RETURN WHENCE U CAME +1
01280 MARK1   LD      A,1             ;MARK1 ONLY FOR FLAG
01290         OUT     (0),A           ;SEND MARK TONE
01300         LD      A,5             ;1 = SPACE & 5 = MARK
01310         LD      (LASONE),A      ;UPDATE LASTONE
01320         XOR     A               ;ZERO OUT 'A' REGISTER
01330         LD      (ZEROSP),A      ;AND ZERO SPACE COUNTER
01340         LD      A,(SPEED)       ;COUNTDOWN VALUE
01350         LD      HL,DECMK1       ;JP (HL) ADDRESS
01360 DECMK1  DEC     A               ;-1 COUNTDOWN VALUE
01370         RET     Z               ;RETURN WHENCE U CAME +1
01380         JP      (HL)            ;JUMP TO DECMK1
01390 SN1A    LD      D,A             ;SN1A ONLY FOR FLAG
01400         LD      E,8             ;NUMBER OF BITS PER BYTE
01410 SN2A    LD      A,(LASONE)      ;1 = SPACE & 5 = MARK
01420         CP      1               ;WAS IT A SPACE ?
01430         JP      Z,LASSP         ;IF SO, GOTO LAST SPACE
01440         BIT     0,D             ;SET Z FLAG FOR BIT ZERO
01450         CALL    NZ,MARK1        ;IF NOT ZERO SEND MARK
01460         BIT     0,D             ;SET Z FLAG FOR BIT ZERO
01470         CALL    Z,SPACE1        ;IF ZERO SEND SPACE
01480         DEC     E               ;-1 FROM BIT COUNTER
01490         RET     Z               ;IF ZERO, GET NEXT BYTE
01500         RRC     D               ;RIGHT SHIFT ALL 1 BIT
01510         JP      SN2A            ;GO BACK FOR NEXT BIT
01520 LASSP   BIT     0,D             ;SET Z FLAG FOR BIT ZERO
01530         CALL    NZ,SPACE1       ;IF NOT ZERO SEND SPACE
01540         BIT     0,D             ;SET Z FLAG FOR BIT ZERO
01550         CALL    Z,MARK1         ;IF ZERO SEND MARK
01560         DEC     E               ;-1 FROM BIT COUNTER
01570         RET     Z               ;IF ZERO, GET NEXT BYTE
01580         RRC     D               ;RIGHT SHIFT ALL 1 BIT
01590         JP      SN2A            ;GO BACK FOR NEXT BIT
01600 ZEROSP  DEFB    0               ;SPACE COUNTER STASH
01610 ZEROMK  DEFB    0               ;MARK COUNTER STASH
01620 SPEED   DEFB    98              ;MODEL 3 USE 115 DECIMAL
01630 ; - - - - - - - - - - - - - - - - - - - - - - - - -
01640 ; END OF FUNDAMENTAL 1200 BAUD TRANSMIT SYNC. CONVERSION
```

entered in line 560 with the first data byte in the 'A' register after the last opening flag was received.

The software digital phase locked loop in lines 980 through 1330 is the author's favorite and is the one used in Volume I of *Synchronous Packet Radio Using The Software Approach.* This DPLL accommodates 1200 baud signals whose timing is off about plus or minus 10 percent from the norm.

software digital phase-locked loop

Figure 3 shows two 1200 baud, bit time frames where the incoming signal has changed from a space to a mark. In SDLC/HDLC it is not the absolute value of a mark or space that counts. It is only the relative change from the previous bit that determines whether it is a logical 1 or logical zero.

The software digital phase-locked loop is divided into quadrants much like its hardware counterpart, the Intel 8273 synchronous data link controller chip. But it is somewhat different in that there is no "dead band" between tyme 2 and tyme 3. If the transition occurs during tyme 2, a bit early from the last PROCESSing, the countdown time delay from tyme 4 is decreased slightly so that the next PROCESSing between tyme 4 and tyme 1 will be closer to dead center. If the transition occurs during tyme 3, a bit late from the last PROCESSing, the countdown time delay for tyme 4 is increased slightly so that the next PROCESSing between tyme 4 and tyme 1 will be closer to dead center. Any transitions during tyme 1 or tyme 4 dramatically decrease or increase, respectively, the countdown time delay to quickly bring the software DPLL back into correct synchronization.

Ideally, the software DPLL countdown values for

fig. 2. Receive mode real-time SDLC/HDLC serial synchronous date stream to parallel decimal byte conversion.

```
00100 ;                    FIGURE 2
00110
00120 ; RECEIVE MODE REAL-TIME SDLC/HDLC SERIAL SYNCHRONOUS
00130 ; DATA STREAM TO PARALLEL DECIMAL BYTE CONVERSION.
00140
00150 ; THE REGISTERS USED IN THIS RECEIVE MODE SUBROUTINE ARE:
00160
00170 ; REGULAR REGISTERS:
00180
00190 ; A = USED + NEW PORT ZERO VALUE IN EACH DPLL QUADRANT
00200 ; F = USED THROUGHOUT
00210 ; B = DPLL COUNTDOWN VALUE FOR FIRST 3 DPLL QUADRANTS
00220 ; C = 8 BITS PER BYTE COUNTER
00230 ; D = CALCULATED DPLL COUNTDOWN VALUE FOR 4TH QUADRANT
00240 ; E = LAST PORT ZERO VALUE
00250 ; HL= MEM LOCATION TO STORE ENDING FLAG ADDRESS
00260 ; IX= ONLY FOR EQUALIZING TIME DELAYS; INC IX & DEC IX
00270 ; IY= UNUSED
00280
00290 ; ALTERNATE REGISTERS:
00300
00310 ; A = UNUSED
00320 ; F = UNUSED
00330 ; B = RECEIVED PARALLEL BYTE WITH ZERO-DELETION
00340 ; C = RECEIVED PARALLEL BYTE WITHOUT ZERO-DELETION
00350 ; D = INCOMING BIT VALUE AT CENTER OF BIT TIME FRAME
00360 ; E = LAST BIT VALUE AT CENTER OF BIT TIME FRAME
00370 ; HL= MEM LOCATION TO STORE CONVERTED DECIMAL BYTE
00380
00390 ; THIS SUBROUTINE IS ENTERED IN LINE 700 WITH THE FIRST
00400 ; ASSEMBLED DATA BYTE IN 'A' REGISTER AFTER THE LAST
00410 ; OPENING FLAG HAS BEEN RECEIVED.
00420
00430 ; THE SOFTWARE DIGITAL PHASE LOCKED LOOP (DPLL) IS AT THE
00440 ; END OF THIS SUBROUTINE IN LINES 980 THROUGH 1330.
00450
00460 IN1     BIT     0,A             ;PACKET TONES DROPPED ?
00470         JP      Z,MOVEM+1       ;IF SO, PROCESS IT.
00480         EXX                     ;SWAP ALTERNATE REGISTERS
00490         LD      D,A             ;INCOMING BIT VALUE TO D
00500         XOR     E               ;COMPARE WITH LAST ONE
00510         LD      E,D             ;UPDATE E FOR NEXT TIME
00520         CPL                     ;DATA IN BIT 7
00530         RLCA                    ;SHIFT INTO CARRY
00540         RR      B               ;OUTPUT DATA -
00550         RRCA                    ;ACCUMULATES HERE.
00560         RR      C               ;INCOMING BIT PATTERN
00570         LD      A,C             ;TEST IT
00580         CP      126             ;FOR A CLOSING FLAG ?
00590         JP      Z,FL1           ;IF SO, GOTO LINE 760
00600         CP      254             ;PACKET TONES DROPPED ?
00610         JP      Z,MOVFM         ;IF SO, PROCESS IT
00620         AND     254             ;REMOVE BIT ZERO
00630         CP      124             ;0111110X PATTERN ?
00640         JP      Z,DELETE        ;IF SO, DO ZERO DELETION
00650         LD      A,B             ;BUILT UP DATA VALUE
00660         EXX                     ;RESTORE REG. REGISTERS
00670         DEC     C               ;DECREMENT BIT COUNTER
00680         JP      NZ,IN4          ;NOT ZERO, GET NEXT BIT
00690 IN1A    NOP                     ;DO NOTHING AT ALL
00700         EXX                     ;SWAP ALTERNATE REGISTERS
00710         INC     HL              ;CONVERTED BYTE LOCATION
00720 IN2     LD      (HL),A          ;STASH IT IN MEMORY
00730         LD      A,H             ;TOO LONG A PACKET ?
00740         CP      176             ;OVER 4096 BYTES LONG ?
00750         JP      Z,MOVEM-3       ;IF SO, PROCESS IT
00760 IN3     EXX                     ;RESTORE REG. REGISTERS
00770         LD      C,8             ;RESET BITS/BYTE COUNTER
00780 IN4     CALL    TYME            ;DIGITAL PHASE LOCK LOOP
00790         JP      IN1             ;CONVERT INCOMING BIT
00800 FL1     PUSH    HL              ;GOT A CLOSING FLAG
00810         EXX                     ;RESTORE REG. REGISTERS
00820         POP     BC              ;FLAG LOCATION MINUS ONE
00830         INC     BC              ;FLAG MEM LOCATION
00840         LD      (HL),C          ;STORE FLAG ADDRESS LSB
00850         INC     HL              ;NEXT STORE LOCATION
00860         LD      (HL),B          ;STORE FLAG ADDRESS MSB
00870         INC     HL              ;NEXT STORE LOCATION
00880         LD      A,144           ;OUT OF BOUNDS DUE TO -
00890         CP      H               ;RUN AWAY TNC ?
00900         JP      Z,MOVEM+1       ;IF SO, PROCESS IT
00910         JP      IN3+1           ;ELSE GO FOR NEXT ONE
00920 DELETE  RL      B               ;ZERO DELETION, SO -
00930         EXX                     ;BACKUP ALTERNATE B
00940         INC     (IX)            ;COMPENSATING TIME -
00950         DEC     (IX)            ;DELAY.
00960         CALL    TYME            ;DIGITAL PHASE LOCK LOOP
00970         JP      IN1             ;CONVERT NEXT BIT
00980 TYME    LD      A,(14400)       ;ESCAPE IS CLEAR KEY
00990         CP      2               ;IF PRESSED GOTO -
01000         JP      Z,MENU0-1       ;TRANSMIT MODE.
01010         LD      B,23            ;MODEL I COUNTDOWN VALUE
01020 TYME1   DJNZ    TYME1           ;1ST QUADRANT COUNTDOWN
01030         IN      A,(0)           ;PORT ZERO VALUE TO 'A'
01040         CP      E               ;ANY CHANGE FROM LAST ?
01050         JP      NZ,DEC2         ;IF SO, GOTO LINE 1180
01060         LD      E,23            ;MODEL I COUNDOWN VALUE
01070 TYME2   DJNZ    TYME2           ;2ND QUADRANT COUNTDOWN
01080         IN      A,(0)           ;PORT ZERO VALUE TO 'A'
01090         CP      E               ;ANY CHANGE FROM LAST ?
01100         JP      NZ,DEC1         ;IF SO, GOTO LINE 1210
01110         LD      B,23            ;MODEL I COUNTDOWN VALUE
01120 TYME3   DJNZ    TYME3           ;3RD QUADRANT COUNTDOWN
01130         IN      A,(0)           ;PORT ZERO VALUE TO 'A'
01140         CP      E               ;ANY CHANGE FROM LAST ?
01150         JP      NZ,INC1         ;IF SO, GOTO LINE 1240
01160         LD      B,D             ;ADJUSTED COUNTDOWN VALUE
01170 TYME4   DJNZ    TYME4           ;4TH QUADRANT COUNTDOWN
01180         IN      A,(0)           ;PORT ZERO VALUE TO 'A'
01190         CP      E               ;ANY CHANGE FROM LAST ?
01200         JP      NZ,INC2         ;IF SO, GOTO LINE 1270
01210         RET                     ;DPLL DONE. GO PROCESS IT
01220 DEC2    LD      E,A             ;SAVE NEW BIT IN 'E'
01230         LD      D,15            ;WAY TOO LATE, SO SHORT-
01240         JP      TYME2-2         ;EN LAST QUADRANT COUNT
01250 DEC1    LD      E,A             ;SAVE NEW BIT IN 'E'
01260         LD      D,20            ;TINY BIT TOO LATE, SO -
01270         JP      TYME3-2         ;SHORTEN LAST QUADRANT.
01280 INC1    LD      E,A             ;SAVE NEW BIT IN 'E'
01290         LD      D,24            ;TINY BIT TOO SOON, SO -
01300         JP      TYME4-2         ;LENGTHEN LAST QUADRANT.
01310 INC2    LD      E,A             ;SAVE NEW BIT IN 'E'
01320         LD      D,29            ;WAY TOO SOON, SO -
01330         RET                     ;LENGTHEN LAST QUADRANT.
01340 ; - - - - - - - - - - - - - - - - - - - - - - - - - - -
01350 ; END OF 1200 BAUD RECEIVE MODE SUBROUTINE
01360
01370 ; IDEALLY THE DPLL WILL OSCILLATE BETWEEN DEC1 & INC1.
01380
01390 ; THE MODEL 3 NORMAL COUNTDOWN VALUE IS 28 FOR B REGISTER
01400 ; IN LINES 1010, 1060, AND 1110. ALL ELSE THE SAME.
```

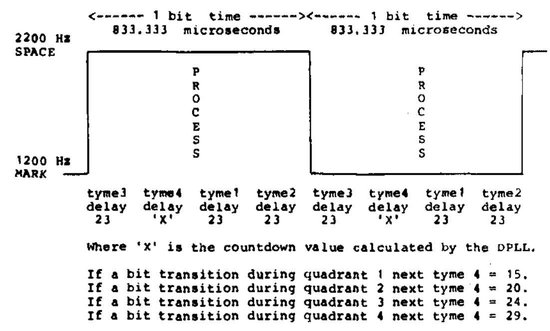

fig. 3. 1200 baud software digital phase-locked loop quadrants.

tyme 4 will oscillate between 20 and 24, which is exactly what this software DPLL accomplishes. With the Model 3's faster clock, register B in TYME, TYME1, and TYME2 is set to 28 decimal.

The PROCESSing time between tyme4 and tyme1 should be held to less than 10 percent of the total 1200 baud 833.333 microsecond bit time so as not to further complicate the software DPLL's job.

virtual real-time CRC generation and checking

Figure 4 is the AX.25 Volume 2 program's CRC subroutine with comments. It uses the 'byte-wise' look-up-table approach suggested by Aram Perez[4] and converted by the author to the CRC polynomial used by both the Vancouver and AX.25 protocols. It is *incredibly* fast. That is why we use the term 'virtual' to describe its almost real-time speed. It is exactly 27 times faster than the bit-by-bit CRC'ing described by the author in Volume 1 of *Synchronous Packet Radio Using the Software Approach*.

Unfortunately, there's no "free lunch." The price we have to pay for this lightning-fast speed is the 512 byte look-up table illustrated at the end of fig. 4. It is shown as a two-byte word table to conserve space. The label TABLE is in the upper left hand corner at

fig. 4. IBM SDLC CRC generation and CR checking subroutines.

```
00100 ;                    FIGURE 4
00110
00120 ; IBM SDLC CRC GENERATION AND CRC CHECKING SUBROUTINES
00130
00140 ; CRC1 AND CRC2 ARE FOR GENERATING THE 2 BYTE CRC VALUE
00150 ; FOR A FRAME OF (LENG1) BYTES IN LENGTH.  ADDRE2 IS THE
00160 ; MEMORY LOCATION OF THE BEGINNING OF THE SINGLE FRAME
00170 ; PACKET TO BE TRANSMITTED.  MULTI-FRAME PACKETS USE A
00180 ; VARIABLE ADDRE2 DEPENDING UPON WHERE EACH FRAME IS
00190 ; SEQUENTIALLY LOCATED IN MEMORY.
00200
00210 ; RCRC BEGINNING IN LINE 800 TESTS THE RECEIVED CRC VALUE
00220 ; OF A FRAME STARTING AT (BGINIT) IN MEMORY WITH A TOTAL
00230 ; LENGTH OF 'BC' REGISTER BYTES.  MULTI-FRAME PACKETS OF
00240 ; 1 TO 7 FRAMES/PACKET ARE ACCOMODATED.
00250
00260 ; TABLE BEGINNING AT LINE 970 IS THE LOOK-UP TABLE FOR
00270 ; THE BRILLIANT 'BYTE WISE' CRC SUBROUTINE SUGGESTED BY
00280 ; ARAM PEREZ IN THE THE JUNE '83 ISSUE OF I.E.E.E. MICRO.
00290
00300 CRCVAL  DEFW    0               ;RECEIVE CRC VALUE STASH
00310 ENDCRC  DEFW    0               ;XMIT CRC VALUE STASH
00320 CRC1    LD      HL,ADDRE2       ;BEGIN MESSAGE LOCATION
00330         LD      BC,(LENG1)      ;LENGTH OF FRAME IN BYTES
00340         LD      DE,65535        ;INTIALIZE DIVIDEND 1'S
00350         CALL    CRCT            ;GENERATE CRC LINE 470
00360         CALL    FINCRC          ;SORT/STUFF RIGHT ORDER
00370         LD      A,(SIGN2)       ;DISPLAY CRC VALUE -
00380         CP      1               ;ON VIDEO DISPLAY ?
00390         RET     Z               ;IF NOT, RETURN.
00400         LD      HL,(ENDCRC)     ;IF SO, THEN DISPLAY IT
00410         CALL    DZ              ;ON TOP LINE OF VIDEO.
00420 CRC2    LD      BC,960          ;= 15 LINES OF VIDEO
00430         LD      HL,ADDREZ       ;BEGIN PACKET ADDRESS
00440         LD      DE,15424        ;2ND LINE OF VIDEO
00450         LDIR                    ;DISPLAY MESSAGE SENT
00460         RET                     ;RETURN WHENCE U CAME +1
00470 CRCT    LD      A,(HL)          ;FIRST BYTE TO CRC
00480         INC     HL              ;INCREMENT FOR NEXT ONE
00490         PUSH    BC              ;SAVE BYTES TO CRC
00500         PUSH    HL              ;SAVE NEXT BYTE LOCATION
00510         XOR     E               ;XOR REMAINDER LSB W/'A'
00520         LD      C,A             ;SAVE RESULT IN 'C'
00530         LD      B,0             ;ZERO OUT 'B'
00540         LD      HL,TABLE        ;LOOKUP TABLE LOCATION
00550         ADD     HL,BC           ;ADD BC TO LOCATION
00560         ADD     HL,BC           ;ADD BC TO LOCATION
00570         LD      A,D             ;REMAINDER MSB TO 'A'
00580         XOR     (HL)            ;XOR WITH TABLE VALUE
00590         LD      E,A             ;SAVE RESULT IN 'E'
00600         INC     HL              ;NEXT TABLE LOCATION
00610         LD      D,(HL)          ;SAVE VALUE IN 'E'
00620         POP     HL              ;NEXT BYTE TO CRC MEM
00630         POP     BC              ;NUMBER BYTES TO CRC
00640         DEC     BC              ;LESS ONE
00650         LD      A,B             ;TEST FOR
00660         OR      C               ;ZERO
00670         JP      NZ,CRCT         ; IF NOT, CRC NEXT ONE
00680         RET                     ;ELSE ALL DONE, RETURN
00690 FINCRC  LD      A,E             ;DE = CRC 2 BYTE VALUE
00700         CPL                     ;COMPLEMENT IT
00710         LD      HL,(WHER4B)     ;END OF MESSAGE +1
00720         LD      (HL),A          ;LD 1ST CRC ON MESSAGE
00730         LD      (ENDCRC+1),A    ;AND SAVE IT HERE
00740         INC     HL              ;NEXT MESSAGE LOCATION
00750         LD      A,D             ;SECOND CRC BYTE
00760         CPL                     ;COMPLEMENT IT
00770         LD      (HL),A          ;LD 2ND CRC ON MESSAGE
00780         LD      (ENDCRC),A      ;AND SAVE IT HERE
00790         RET                     ;RETURN WHENCE U CAME +1
00800
00810 RCRC    LD      DE,65535        ;RECEIVE CRC CHECK
00820         LD      HL,(BGINIT)     ;BEGIN FRAME LOCATION
00830         CALL    CRCT            ;CRC ALL INCLUDING CRC
00840         LD      (CRCVAL),DE     ;SAVE REMAINDER IN MEM
00850         LD      HL,61624        ;COMPARE REMAINDER WITH
00860         RST     16H             ;61624 DECIMAL
00870         JP      NZ,BADCRC       ;NOT ZERO = BAD ONE
00880         RET                     ;OK, SO RETURN
00890 BADCRC  POP     AF              ;ADJUST STACK
00900         POP     AF              ;FOR 2 CALLS
00910         LD      IY,37492        ;<BAD CRC> MESSAGE
00920         CALL    SHOWIT          ;DISPLAY ON VIDEO
00930         JP      MODE1A          ;GO AWAIT NEXT PACKET
00940
00950 ; 512 BYTE LOOKUP TABLE FOR BYTE-WISE CRC SUBROUTINE
00960
00970 TABLE   DEFW    0
00980 ; - - - - - - - - - - - - - - - - - - - - - - - - - -
00990
01000 ; SEE CRC LOOK-UP TABLE BELOW
```

```
FIGURE 4 CONTINUED
- - - - - - - - - - - - - - - - - - - - - - - - - - - -

This is the 512 byte CRC lookup table printed out as 256 two
byte words to save space.  The label TABLE is at location 1.

 1 DEFW     0   53 DEFW 30631  105 DEFW 61262  157 DEFW 24293  209 DEFW 54925
 2 DEFW  4489   54 DEFW 26158  106 DEFW 65223  158 DEFW 20332  210 DEFW 50948
 3 DEFW  8978   55 DEFW 21685  107 DEFW 52316  159 DEFW 32247  211 DEFW 62879
 4 DEFW 12955   56 DEFW 17724  108 DEFW 56789  160 DEFW 27774  212 DEFW 58390
 5 DEFW 17956   57 DEFW 48587  109 DEFW 43370  161 DEFW 42250  213 DEFW 37033
 6 DEFW 22445   58 DEFW 44098  110 DEFW 47331  162 DEFW 46211  214 DEFW 33056
 7 DEFW 25910   59 DEFW 40665  111 DEFW 35448  163 DEFW 34328  215 DEFW 46011
 8 DEFW 29887   60 DEFW 36688  112 DEFW 39921  164 DEFW 38801  216 DEFW 41522
 9 DEFW 35912   61 DEFW 64495  113 DEFW 29575  165 DEFW 58158  217 DEFW 23237
10 DEFW 40385   62 DEFW 60006  114 DEFW 25102  166 DEFW 62119  218 DEFW 19276
11 DEFW 44890   63 DEFW 55549  115 DEFW 20629  167 DEFW 49212  219 DEFW 31191
12 DEFW 48851   64 DEFW 51572  116 DEFW 16668  168 DEFW 53685  220 DEFW 26718
13 DEFW 51820   65 DEFW 16900  117 DEFW 13731  169 DEFW 10562  221 DEFW  7393
14 DEFW 56293   66 DEFW 21389  118 DEFW  9258  170 DEFW 14539  222 DEFW  3432
15 DEFW 59774   67 DEFW 24854  119 DEFW  5809  171 DEFW  2640  223 DEFW 16371
16 DEFW 63735   68 DEFW 28831  120 DEFW  1848  172 DEFW  7129  224 DEFW 11898
17 DEFW  4225   69 DEFW  1056  121 DEFW 65487  173 DEFW 28518  225 DEFW 59150
18 DEFW   264   70 DEFW  5545  122 DEFW 60998  174 DEFW 32495  226 DEFW 63111
19 DEFW 13203   71 DEFW 10034  123 DEFW 56541  175 DEFW 19572  227 DEFW 50204
20 DEFW  8730   72 DEFW 14011  124 DEFW 52564  176 DEFW 24061  228 DEFW 54677
21 DEFW 22181   73 DEFW 52812  125 DEFW 47595  177 DEFW 46475  229 DEFW 41258
22 DEFW 18220   74 DEFW 57285  126 DEFW 43106  178 DEFW 41986  230 DEFW 45219
23 DEFW 30135   75 DEFW 60766  127 DEFW 39673  179 DEFW 38553  231 DEFW 33336
24 DEFW 25662   76 DEFW 64727  128 DEFW 35696  180 DEFW 34576  232 DEFW 37809
25 DEFW 40137   77 DEFW 34920  129 DEFW 33800  181 DEFW 62383  233 DEFW 27462
26 DEFW 36160   78 DEFW 39393  130 DEFW 38273  182 DEFW 57894  234 DEFW 31439
27 DEFW 49115   79 DEFW 43898  131 DEFW 42778  183 DEFW 53437  235 DEFW 18516
28 DEFW 44626   80 DEFW 47859  132 DEFW 46739  184 DEFW 49460  236 DEFW 23005
29 DEFW 56045   81 DEFW 21125  133 DEFW 49708  185 DEFW 14787  237 DEFW 11618
30 DEFW 52068   82 DEFW 17164  134 DEFW 54181  186 DEFW 10314  238 DEFW 15595
31 DEFW 63999   83 DEFW 29079  135 DEFW 57662  187 DEFW  6865  239 DEFW  3696
32 DEFW 59510   84 DEFW 24606  136 DEFW 61623  188 DEFW  2904  240 DEFW  8185
33 DEFW  8450   85 DEFW  5281  137 DEFW  2112  189 DEFW 32743  241 DEFW 63375
34 DEFW 12427   86 DEFW  1320  138 DEFW  6601  190 DEFW 28270  242 DEFW 58886
35 DEFW   528   87 DEFW 14259  139 DEFW 11090  191 DEFW 23797  243 DEFW 54429
36 DEFW  5017   88 DEFW  9786  140 DEFW 15067  192 DEFW 19836  244 DEFW 50452
37 DEFW 26406   89 DEFW 57037  141 DEFW 20068  193 DEFW 50700  245 DEFW 45483
38 DEFW 30383   90 DEFW 53060  142 DEFW 24557  194 DEFW 55173  246 DEFW 40994
39 DEFW 17460   91 DEFW 64991  143 DEFW 28022  195 DEFW 58654  247 DEFW 37561
40 DEFW 21949   92 DEFW 60502  144 DEFW 31999  196 DEFW 62615  248 DEFW 33584
41 DEFW 44362   93 DEFW 39145  145 DEFW 38025  197 DEFW 32808  249 DEFW 31687
42 DEFW 48323   94 DEFW 35168  146 DEFW 34048  198 DEFW 37281  250 DEFW 27214
43 DEFW 36440   95 DEFW 48123  147 DEFW 47003  199 DEFW 41786  251 DEFW 22741
44 DEFW 40913   96 DEFW 43634  148 DEFW 42514  200 DEFW 45747  252 DEFW 18780
45 DEFW 60270   97 DEFW 25350  149 DEFW 53933  201 DEFW 19012  253 DEFW 15843
46 DEFW 64231   98 DEFW 29327  150 DEFW 49956  202 DEFW 23501  254 DEFW 11370
47 DEFW 51324   99 DEFW 16404  151 DEFW 61887  203 DEFW 26966  255 DEFW  7921
48 DEFW 55797  100 DEFW 20893  152 DEFW 57398  204 DEFW 30943  256 DEFW  3960
49 DEFW 12675  101 DEFW  9506  153 DEFW  6337  205 DEFW  3168
50 DEFW  8202  102 DEFW 13483  154 DEFW  2376  206 DEFW  7657
51 DEFW  4753  103 DEFW  1584  155 DEFW 15315  207 DEFW 12146
52 DEFW   792  104 DEFW  6073  156 DEFW 10842  208 DEFW 16123
```

location 1. The table resides sequentially in memory from location 1 through 256 anywhere you wish to put it.

adapting the software approach to non-TRS-80 computers

This is neither a simple nor impossible task if you are a professional assembly language programmer with access to a mini or mainframe computer with cross-assemblers for the more popular microprocessors. Many of the software houses now use this accessory. The alternative is to create your own software approach packet program for your own favorite microcomputer. Hopefully this short article will inspire you to do so.

conclusion

Further information on the software approach can be found in *Synchronous Packet Radio Using The Software Approach — Volume II: AX.25 Protocol.*[3] This 253-page book is available for $22 postpaid (overseas, add $10 for airmail) from Richcraft Engineering Ltd., #1 Wahmeda Industrial Park, Chautauqua, New York 14722, or from Ham Radio's Bookstore, Greenville, N. H. 03048.

If you wish to forego the pleasure of typing in approximately 5000 lines of source code, the program is available (from Richcraft) on disk for the Model I or Model III TRS-80 (specify) for an additional $29, postpaid. If you wish to modify the program with your call letters or personalized prepared messages, you'll need a copy of the book, but no knowledge of assembly language, and no editor/assembler.

references

1. Margaret Morrison, KV7D, and Dan Morrison, KV7B, "Amateur Packet Radio," Part 1, *ham radio*, July 1983, page 14; Part 2, *ham radio*, August, 1983, page 18.
2. Lyle Johnson, WA7GXD, "Join the Packet Radio Revolution," Part 1, *73*, September, 1983, page 19; Part 2, *73*, October, 1983, page 20.
3. Robert M. Richardson, W4UCH, *Synchronous Packet Radio Using The Software Approach*, Volumes I - III, Richcraft Engineering, Ltd., 1982.
4. Aram Perez, "Byte-wise CRC Calculations," *IEEE Micro Journal*, IEEE Computer Society (10662 Los Vaqueros Circle, Los Alamitos, California 90720), June, 1983, pages 41.50.

ham radio

interrupt-driven RTTY reader

Receive RTTY on your Apple while other programs run

This is a 6502 machine language program written for an Apple II equipped with a 6522 Versatile Interface Adapter chip (VIA).* (It should be simple to convert the program to other 6502 computers.) The unusual feature of this program is that RTTY reception is accommodated during interrupts that occur at 1000 Hz intervals. Thus, this program can easily be extended without concern about timing. This program will not work well on the Apple IIe at 100 WPM because when the IIe does a scrolling operation, it turns off the interrupts! This is because the IIe allows the alternate 64K to bank-select over the ROMs, which contain the IRQ vector. (This confirms my hunch that Apple did not really have interrupts in mind when they designed the IIe.)

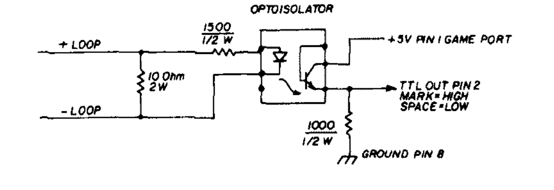

fig. 1. 170 volt current loop to TTL connector.

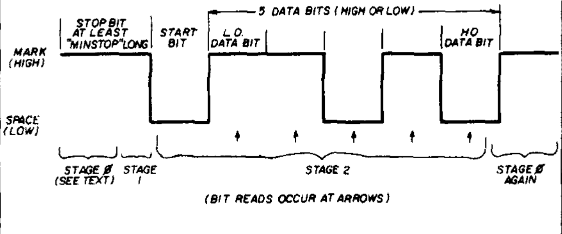

fig. 2. Program operation on a typical character.

*The 6522 is found on some accessory cards (John Bell and Mockingboard, for example) of which several thousand have been sold — **Editor**.

A suitable terminal unit (TU) must be used between the radio receiver and the Apple's game port. The input is switch 0 (pin 2 of the game port). Five volts, or TTL high, should be applied when the TU detects mark condition and zero volts, or TTL low when a space is detected. This is the same input that is commonly used with other programs available for the Apple. Many late-model terminal units have TTL compatible outputs; **fig. 1** shows a simple circuit to convert from current loop to TTL. Be careful with this circuit . . . you don't want to hook up the 170-volt loop to your Apple or to yourself! The circuit uses an optoisolator that should protect your Apple from the high voltages.

The program listing is heavily commented. Here's how it works (see **fig. 2**):

STAGE 0. The program searches for a stop pulse, which is an uninterrupted mark at least as long as "MINSTOP."

STAGE 1. After finding a stop pulse, the program waits for a transition to space indicating a start pulse.

STAGE 2. The program reads in 5 data bits, waiting the proper amount of time between each bit.

Program operation is simple. Just assemble the program and 'BRUN' it from the disk. Values for different speeds are included in the listing. It supports 80-column display boards and printers as well; just type "PR#3" or "PR#1" to turn on the board or printer before running the program.

Many enhancements are possible — for example, creating a larger text buffer so that the main program can do other tasks while still receiving RTTY, or for later transfer of the received text to disk or printer. Some non-ham TTY stations broadcast using inversion of one or more bits of the five-level code; this could be decoded by "exclusive-or" with various values until readable copy appears. It should be possible to add enhancements in basic because interrupts are used. Of course, a transmit function could be added, and this program could even form the basis for a powerful "MSO," or bulletin board.

By Scott D. Schram, KN4L, 225 LaPrado Place, Homewood, Alabama 35209.

```
                   ********************************
                   *                              *
                   *     RTTY READER PROGRAM      *
                   *      BY SCOTT D. SCHRAM      *
                   *                              *
                   ********************************
                   *
                            ORG  $803
                   *
                   *    *** EQUATES ***
                   *
                   * RECIEVED CHARACTER BUFFER.
                   *
                   RECCHAR  =    $4000
                   *
                   COUT     =    $FDED       ;APPLE'S OUTPUT CHAR IN 'A' REG
                   IRQVEC   =    $3FE        ;APPLE'S IRQ VECTOR LOCATION
                   *
                   * RTTY INPUT IS APPLE SWITCH 0
                   *  MARK IS HIGH, MAKING RTTYIN NEGATIVE
                   *  SPACE IS LOW, MAKING RTTYIN POSITIVE
                   *
                   RTTYIN   =    $C061
                   *
                   * THE 6522 VERSATILE INTERFACE ADAPTER LOCATION
                   * (YOU NEED A VIA TO MAKE THIS PROGRAM WORK,
                   *  I USE THE JOHN BELL ENGINEERING BOARD.)
                   *  SOME OTHER PERIPHERAL CARDS HAVE A VIA AVAILABLE.
                   *  SOME OTHER COMPUTERS HAVE A VIA BUILT IN.
                   *
                   **** THE VIA MUST BE CONNECTED TO THE IRQ LINE
                   ****  SO IT CAN INTERRUPT THE PROCESSOR
                   *
                   * CHANGE THE VIA EQUATE TO YOUR VIA LOCATION
                   *
                   VIA      =    $C200       ;SLOT 2 VIA BASE LOCATION
                   ACR      =    VIA+11      ;VIA AUXILLIARY CONTROL REG.
                   IER      =    VIA+14      ;VIA INTERRUPT ENABLE REG.
                   IFR      =    VIA+13      ;VIA INTERRUPT FLAG REG.
                   T1L      =    VIA+4       ;VIA TIMER 1 LOW BYTE
                   T1H      =    VIA+5       ;VIA TIMER 1 HIGH BYTE
                   *
                   ********************************
                   *
                   *    *** MAIN FOREGROUND ROUTINES ***
                   *
0803: A9 00        START    LDA  #0          ;CLEAR RECIEVED CHARACTER
0805: 8D 49 08              STA  RECPNTA
0808: 8D 4A 08              STA  RECPNTB
                   *
080B: A9 00                 LDA  #0
080D: 8D 4B 08              STA  STAGE       ;INIT INTERR. VARS.
0810: 8D 4C 08              STA  MARKCNT
0813: 8D 55 08              STA  SHIFT
                   *
0816: A9 5C                 LDA  #<INTSVC    ;SETUP APPLE'S IRQ VECTOR
0818: 8D FE 03              STA  IRQVEC
081B: A9 08                 LDA  #>INTSVC
081D: 8D FF 03              STA  IRQVEC+1
                   *
                   * SETUP THE VIA REGISTERS TO INTERRUPT
                   *  1000 TIMES PER SECOND.
                   *
0820: A9 40                 LDA  #%01000000
0822: 8D 0B C2              STA  ACR         ;TIMER 1 FREE RUNNING
0825: A9 FF                 LDA  #255
0827: 8D 04 C2              STA  T1L
082A: A9 03                 LDA  #3
082C: 8D 05 C2              STA  T1H         ;1000 HZ INTERRUPTS
082F: A9 C0                 LDA  #%11000000
0831: 8D 0E C2              STA  IER         ;ENABLE VIA INTERRUPTS
0834: 58                    CLI              ;ENABLE IRQ
                   *
                   * WAIT FOR CHARACTER TO BE RECIEVED BY INTERRUPT ROUTINE
                   *
0835: AE 49 08     WAIT     LDX  RECPNTA
0838: EC 4A 08              CPX  RECPNTB
083B: F0 F8                 BEQ  WAIT        ;NONE YET.
083D: EE 49 08              INC  RECPNTA     ;MOVE PRINT POINTER UP ONE
0840: BD 00 40              LDA  RECCHAR,X   ;GET CHARACTER
0843: 20 ED FD              JSR  COUT        ;PRINT IT.
0846: 4C 35 08              JMP  WAIT        ;WAIT FOR MORE
                   *
                   ********************************
                   *
                   *    ***  VARIABLES ***
                   *
                   * IF THESE TWO POINTERS ARE UNEQUAL THEN
                   * AT LEAST ONE CHARACTER IS AVAILABLE FOR PRINTING
                   *
                   RECPNTA  DS   1
                   RECPNTB  DS   1
                   *
                   * STAGE=0, WAITING FOR STOP PULSE
                   * STAGE=1, FOUND STOP PULSE, NOW WAIT FOR START PULSE
                   * STAGE=2, READING SUCCESSIVE BITS
                   *
                   STAGE    DS   1
                   *
                   * MARKCNT IS USED WHEN STAGE=0 TO COUNT UP
                   *  DURATION OF THE STOP PULSE (WHICH IS A MARK)
                   *  AND SEE IF IT IS LONG ENOUGH.
                   *
                   MARKCNT  DS   1
                   *
                   * BITCNT IS USED WHEN STAGE=2 TO TELL WHICH BIT
                   *  IS TO BE READ NEXT. (RANGES FROM 0 TO 4)
                   *
                   BITCNT   DS   1
                   *
                   * BITWAIT IS USED WHEN STAGE=2 TO COUNT DOWN DELAYS
                   *  BETWEEN SUCCESSIVE BITS.
                   *
                   BITWAIT  DS   1
                   *
                   * BITS STORES THE 5 ACTUAL BIT VALUES READ DURING
                   *  STAGE TWO FOR ANALYSIS WHEN THE ENTIRE
                   *  CHARACTER HAS BEEN RECIEVED.
                   *
                   BITS     DS   5           ;BIT VALUES READ (L.O. FIRST)
                   *
                   * CHAR IS USED AS TEMPORARY WORK SPACE TO
                   *  BUILD UP THE BAUDOT VALUE FROM THE
                   *  INDIVIDUAL BITS READ.
                   *
                   CHAR     DS   1
                   *
                   * SHIFT SHOWS IF THE MACHINE IS RECIEVING LETTERS
                   *  OR NUMBERS. (AND WHICH TABLE TO LOOKUP IN.)
                   *      0 = LETTERS.
                   *  NON-0 = NUMBERS.
                   *
                   SHIFT    DS   1
                   *
                   *    *** CHANGE MINSTOP AND DELAYS FOR SPEED CHANGE ***
                   *
                   * MINIMUM ACCEPTABLE STOP PULSE LENGTH
                   *  (MUST REMAIN MARK FOR AT LEAST THIS LONG
                   *   TO BEGIN READING A CHARACTER)
                   *
                   * MINSTOP=30 FOR 60 WPM.
                   *         27 FOR 66 WPM.
                   *         16 FOR 100 WPM.
                   *
0856: 10           MINSTOP  DFB  16          ;(100 WPM)
                   *
                   * DELAYS FROM THE BEGINNING OF THE START PULSE
                   *  TO THE MIDDLE OF EACH DATA BIT.
                   *
                   * DELAYS ARE 33,22,22,22,22 FOR 60 WPM.
                   *            30,20,20,20,20 FOR 66 WPM.
                   *            20,15,14,10,14 FOR 100 WPM.
                   *
0857: 14 0D 0E     DELAYS   DFB  20,18,14,13,14 ;(100 WPM)
085A: 0D 0E
                   *
                   ********************************
                   *
                   *     *** INTERRUPT SERVICE ROUTINE ***
                   *
085C: 78           INTSVC   SEI              ;DISABLE IRQ
                   *
                   * SAVE REGISTERS
                   *  ('A' REGISTER WAS SAVED AT $45 BY APPLE MONITOR.
                   *   IF YOU USE INTERRUPTS DON'T USE $45!!!)
                   *
085D: 98                    TYA
085E: 48                    PHA              ;SAVE Y
085F: 8A                    TXA
0860: 48                    PHA              ;SAVE X
                   *
0861: 2C 04 C2              BIT  T1L         ;CLEAR TIMER INTERRUPT FLAG
                   *
0864: AD 4B 08              LDA  STAGE
0867: D0 1E                 BNE  INTSVC1
                   *
                   * WAITING FOR STOP PULSE
                   *
0869: AD 61 C0              LDA  RTTYIN
086C: 30 03                 BMI  IS0A        ;MARK?
086E: 4C C7 08              JMP  NEWCHAR     ;ENCOUNTERED SPACE TOO SOON
0871: EE 4C 08     IS0A     INC  MARKCNT     ;MARK RECIEVED
0874: AD 4C 08              LDA  MARKCNT
0877: CD 56 08              CMP  MINSTOP     ;MINIMUM STOP PULSE LENGTH
087A: B0 03                 BGE  IS0B
087C: 4C CF 08              JMP  EXIT        ;NOT LONG ENOUGH YET
                   *
087F: A9 01        IS0B     LDA  #1          ;FOUND STOP PULSE,
0881: 8D 4B 08              STA  STAGE       ;NOW WAIT FOR START
0884: 4C CF 08              JMP  EXIT
                   *
0887: C9 01        INTSVC1  CMP  #1          ;SEE IF WAITING FOR START PULSE
0889: D0 15                 BNE  INTSVC2     ;NO?
088B: AD 61 C0              LDA  RTTYIN
088E: 10 03                 BPL  IS1A        ;SPACE FOUND!
0890: 4C CF 08              JMP  EXIT        ;NO START YET
0893: A9 02        IS1A     LDA  #2          ;BEGIN WAITING FOR DATA BITS
0895: 8D 4B 08              STA  STAGE
0898: A9 00                 LDA  #0
089A: 8D 4D 08              STA  BITCNT
089D: 4C BB 08              JMP  DELSET
                   *
08A0: CE 4E 08     INTSVC2  DEC  BITWAIT     ;WAIT FOR NEXT DATA BIT
08A3: D0 2A                 BNE  EXIT
08A5: AD 61 C0              LDA  RTTYIN      ;IT'S TIME, SO GET A BIT
08A8: AE 4D 08              LDX  BITCNT
08AB: 9D 4F 08              STA  BITS,X      ;AND SAVE IT FOR LATER ANALYSIS
08AE: EE 4D 08              INC  BITCNT
08B1: AD 4D 08              LDA  BITCNT
08B4: C9 05                 CMP  #5          ;SEE IF WE HAVE GOTTEN 5
08B6: 90 03                 BLT  DELSET      ;NO, SET NEXT DELAY
08B8: 4C D7 08              JMP  FNDCHAR     ;YES, FOUND CHARACTER.
                   *
08BB: AE 4D 08     DELSET   LDX  BITCNT
08BE: BD 57 08              LDA  DELAYS,X    ;GET DELAY TIME TO NEXT BIT
08C1: 8D 4E 08              STA  BITWAIT     ;AND PUT IN COUNTDOWN
08C4: 4C CF 08              JMP  EXIT
                   *
                   *
08C7: A9 00        NEWCHAR  LDA  #0          ;START READING A NEW CHAR
08C9: 8D 4B 08              STA  STAGE       ;PREPARE TO WAIT FOR STOP BIT
08CC: 8D 4C 08              STA  MARKCNT
                   *
                   * *** EXIT INTERRUPT ROUTINE ***
                   *
                   * RESTORE REGISTERS
                   *
08CF: 68           EXIT     PLA
08D0: AA                    TAX              ;RESTORE X
08D1: 68                    PLA
08D2: A8                    TAY              ;RESTORE Y
08D3: A5 45                 LDA  $45         ;RESTORE A
08D5: 58                    CLI              ;RE-ENABLE IRQ
08D6: 40                    RTI              ;RETURN TO MAIN PROGRAM
                   *
                   *    *** ALL 5 BITS READ, PROCESS CHARACTER ***
                   *     (SUBROUTINE USED BY INTERRUPT PROCESSING)
                   *
                   * PUT THE BITS TOGETHER INTO A BAUDOT CHARACTER
                   *
08D7: A9 00        FNDCHAR  LDA  #0          ;CLEAR CHARACTER
08D9: 8D 54 08              STA  CHAR
08DC: A2 04                 LDX  #4          ;DO ALL 5 BITS
08DE: BD 4F 08     FC0      LDA  BITS,X
08E1: 0A                    ASL              ;ROLL BIT INTO CARRY
08E2: 2E 54 08              ROL  CHAR        ;THEN OUT INTO CHARACTER
08E5: CA                    DEX
08E6: 10 F6                 BPL  FC0         ;LOOP FOR ALL 5
                   *
                   * NOW, HAVE THE BAUDOT REPRESENTATION OF THE CHARACTER
                   *  DETERMINE WHAT TO DO WITH IT.
                   *
08E8: AE 54 08              LDX  CHAR
08EB: D0 03                 BNE  FC1         ;TEST FOR 'BLANK'
08ED: 4C C7 08              JMP  NEWCHAR     ;IGNORE BLANK, JUST EXIT.
08F0: E0 08        FC1      CPX  #8          ;TEST FOR CARRIAGE RETURN
08F2: F0 0E                 BEQ  UNSHIFT     ;IGNORE CR BUT UNSHIFT
08F4: E0 1B                 CPX  #27         ;TEST FOR FIGURES
08F6: D0 06                 BNE  FC3
08F8: 8E 55 08              STX  SHIFT       ;SHIFT TO FIGURES
08FB: 4C C7 08              JMP  NEWCHAR     ;AND EXIT
08FE: E0 1F        FC3      CPX  #31         ;TEST FOR LETTERS
0900: D0 08                 BNE  FC4A
0902: A9 00        UNSHIFT  LDA  #0
0904: 8D 55 08              STA  SHIFT       ;LETTERS
0907: 4C C7 08              JMP  NEWCHAR
                   *
                   * UNSHIFT ON LF OR SPACE, THEN PRINT THEM
                   *
090A: E0 02        FC4A     CPX  #2          ;LF
090C: F0 04                 BEQ  FC4B
090E: E0 04                 CPX  #4          ;SPACE
0910: D0 05                 BNE  FC4
0912: A9 00        FC4B     LDA  #0
0914: 8D 55 08              STA  SHIFT       ;UNSHIFT
                   *
                   * GET THE ASCII CHARACTER TO PRINT FROM THE
                   *  APPROPRIATE TABLE DEPENDING ON SHIFT.
                   *
0917: AD 55 08     FC4      LDA  SHIFT
091A: D0 06                 BNE  FC5
091C: BD 31 09              LDA  LTRSTBL,X   ;IF LETTERS SHIFT
091F: 4C 25 09              JMP  FC6
0922: BD 51 09     FC5      LDA  FIGSTBL,X   ;IF FIGURES SHIFT
                   *
0925: AE 4A 08     FC6      LDX  RECPNTB
0928: 9D 00 40              STA  RECCHAR,X   ;STORE CHARACTER IN BUFFER
092B: EE 4A 08              INC  RECPNTB
092E: 4C C7 08              JMP  NEWCHAR
                   *
```

```
                 *   *** THE TABLES OF ASCII VALUES FOR BAUDOT CHARS. ***
                 *
0931: A0 C5      LTRSTBL  ASC  " E"
0933: 8D                  HEX  8D         ;LF
0934: C1 A0               ASC  "A "
0936: D3 C9 D5            ASC  "SIU D"
0939: A0 C4
093B: D2 CA CE            ASC  "RJNFC"
093E: C6 C3
0940: CB D4 DA            ASC  "KTZLW"
0943: CC D7
0945: C8 D9 D0            ASC  "HYPQO"
0948: D1 CF
094A: C2 C7 A0            ASC  "BG MX"
094D: CD D8
094F: D6 A0               ASC  "V "
                 *
0951: A0 B3      FIGSTBL  ASC  " 3"
0953: 8D                  HEX  8D
0954: AD A0               ASC  "- "
0956: 87                  HEX  87         ;BELL
0957: B8 B7 A0            ASC  "87 $4"
095A: A4 B4
095C: A7 AC A1            ASC  "',!:("
095F: BA A8
0961: B5                  ASC  "5"
0962: A2                  HEX  A2         ;DOUBLE QUOTE MARK
0963: A9 B2 A3            ASC  ")2#"
0966: B6 B0 B1            ASC  "6019?"
0969: B9 BF
096B: A6 A0 AE            ASC  "& ./; "
096E: AF BB A0
                 *
                 *********************************

--End assembly--

366 bytes

Errors: 0

Symbol table - alphabetical order:

    ACR      =$C20B     BITCNT   =$084D     BITS     =$084F     BITWAIT  =$084E
    CHAR     =$0854     COUT     =$FDED     DELAYS   =$0857     DELSET   =$088B
    EXIT     =$08CF     FC0      =$08DE     FC1      =$08F0     FC3      =$08FE
    FC4      =$0917     FC4A     =$090A     FC4B     =$0912     FC5      =$0922
    FC6      =$0925     FIGSTBL  =$0951     FNDCHAR  =$08D7     IER      =$C20E
?   IFR      =$C20D     INTSVC   =$085C     INTSVC1  =$0887     INTSVC2  =$08A0
    IRQVEC   =$03FE     IS0A     =$0871     IS0B     =$087F     IS1A     =$0893
    LTRSTBL  =$0931     MARKCNT  =$084C     MINSTOP  =$0856     NEWCHAR  =$08C7
    RECCHAR  =$4000     RECPNTA  =$0849     RECPNTB  =$084A     RTTYIN   =$C061
    SHIFT    =$0855     STAGE    =$0848  ?  START    =$0803     T1H      =$C205
    T1L      =$C204     UNSHIFT  =$0902     VIA      =$C200     WAIT     =$0835

Symbol table - numerical order:

    IRQVEC   =$03FE  ?  START    =$0803     WAIT     =$0835     RECPNTA  =$0849
    RECPNTB  =$084A     STAGE    =$084B     MARKCNT  =$084C     BITCNT   =$084D
    BITWAIT  =$084E     BITS     =$084F     CHAR     =$0854     SHIFT    =$0855
    MINSTOP  =$0856     DELAYS   =$0857     INTSVC   =$085C     IS0A     =$0871
    IS0B     =$087F     INTSVC1  =$0887     IS1A     =$0893     INTSVC2  =$08A0
    DELSET   =$088B     NEWCHAR  =$08C7     EXIT     =$08CF     FNDCHAR  =$08D7
    FC0      =$08DE     FC1      =$08F0     FC3      =$08FE     UNSHIFT  =$0902
    FC4A     =$090A     FC4B     =$0912     FC4      =$0917     FC5      =$0922
    FC6      =$0925     LTRSTBL  =$0931     FIGSTBL  =$0951     RECCHAR  =$4000
    RTTYIN   =$C061     VIA      =$C200     T1L      =$C204     T1H      =$C205
    ACR      =$C20B  ?  IFR      =$C20D     IER      =$C20E     COUT     =$FDED
```

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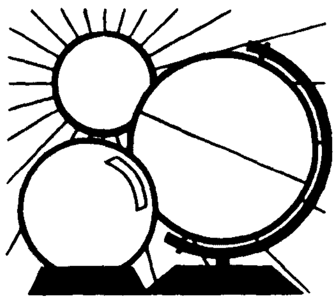


DX FORECASTER

Garth Stonehocker, KØRYW

September is a special season of the year for propagation, and therefore DX. The reason for this is that during the equinox, with its nights and days of equal length, the sun is directly overhead at noon on the geographic equator. This causes solar radiation to hit the earth broadside, and because the equatorial planes of the sun and earth nearly coincide at this time, particles from the sun's eruptions (flares) and coronal holes (thin places in the sun's gases) have a bull's-eye path to the earth. These charged particles, called the solar wind, enter the earth's atmosphere in the polar regions; they also build up in the Van Allen belts around the earth above the equatorial region. When full, the belts release these particles into the polar auroral zone, (on the Canadian-U.S. side after about 2200 local time), causing geomagnetic storms.

Geomagnetic storms affect propagation — and DX — in three ways. First, the particles entering the auroral zone ionospheric D and E regions increase signal absorption, resulting in weak east-west path signals and few signals across the poles. Second, the particles form a reflective curtain along the equatorial side of the auroral zone (south side for us in North America), enhancing VHF auroral scatter propagation. Third, the F region of the ionosphere, once again looking at the auroral zone, but further south, becomes depleted of electrons, forms an electron density trough in which the maximum usable frequency (MUF) for a particular path *through* this area decreases by 30 to 50 percent.

However, still further south at ±20 degrees from the geomagnetic equator, an equivalent-size enhancement of the F region occurs, resulting in evening TE (Trans-Equatorial) openings during the equinox and winter seasons. These three effects vary with time on a short to long basis (seconds through hours), causing what we experience as fading. These effects continue to occur each night for two to three days before ionospheric equilibrium is established. The closer to the equator these effects occur, the bigger the geomagnetic storm (higher K or A value).

Just as the particle density and speed of the solar wind vary, so do the characteristics of the geomagnetic field and ionosphere. Ionospheric variation causes signal focusing and defocusing, which simply means that the signals arriving at your QTH will vary in both strength and angle of arrival. Some directions and locations you haven't heard from in a long time may suddenly be workable, but this kind of surprise is what you can expect during the equinoxes.

last-minute forecast

The higher HF bands, 10 through 30 meters, are expected to be best just after the middle of the month as a result of high solar flux and activity on the sun. Some solar flares that may cause the earth's geomagnetic field to be disturbed for two to three days are possible. However, most of these disturbances will probably be the result of solar coronal hole activity during the extended periods of low solar flux during the first and second weeks of the month. These disturbances will occur on or near September 1, 5, 10, 14, 24, and 30. The highly disturbed period of the sunspot cycle will still be felt in 1984 and 1985. (The first peak of these disturbances occurred in September, 1982; a later peak of this cycle may occur this fall, (in September or October.) Despite these disturbances, DX on the lower frequency bands should be better than it was over the summer months, especially during the second week of September.

The full moon will be visible on the 9th and be at perigee on the 23rd. The autumnal equinox will occur on September 22nd at 2033 UT.

band-by-band summary

Ten and fifteen meters will provide many short-skip E_s openings and long skip openings during the high solar flux periods to most areas of the world during daylight. Some trans-equatorial openings associated with disturbed ionospheric conditions may occur in the evening hours.

Twenty, thirty, and forty meters will support propagation from most areas of the world during the daytime and into the evening hours almost every day, either long-skip to 2500 miles (4000 km) or short-skip E_s to 1250 miles (2000 km) per hop.

Thirty, forty, eighty, and one-sixty meters are all good for nighttime DX. However, on many nights 30 and 40 meters will be the only usable bands because of the effect of thunderstorm QRN on 80 and 160 meters. Signal strength via short-skip E_s may overcome the static when E_s is available, even though E_s propagation does become more scarce in September.

ham radio

SEPTEMBER

WESTERN USA

GMT	PDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	5:00	20	20	15	10	15	10	10	20
0100	6:00	20	20	15	10	15	10	10	20
0200	7:00	20	20	20	10	15	10	10	20
0300	8:00	20	20	20	10	15	10	10	20
0400	9:00	20	20	20	10	20	15	10	20
0500	10:00	20	30	20	10	20	15	15	20
0600	11:00	20	30	20	10	20	15	15	20
0700	12:00	20	*30*	20	15	*20*	20	15	20
0800	1:00	*20*	*30*	*20*	15	*20*	*20*	*20*	20
0900	2:00	*20*	*30*	*20*	*15*	*30*	*20*	*20*	30
1000	3:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	*30*
1100	4:00	*30**	30	*20*	*20*	*30*	*20*	*20*	*30*
1200	5:00	*30*	30	20	20	*30*	*20*	*20*	*30*
1300	6:00	*30*	30	15	20	*30**	*20*	*20*	*30*
1400	7:00	30	20	15	20	*20*	*20*	*20*	*30*
1500	8:00	30	20	15	20	*20*	*20*	*20*	*30*
1600	9:00	30	20	15	20	20	15	*20*	*30*
1700	10:00	30	20	15	20	20	15	20	*30*
1800	11:00	30	20	15*	20	20	15	15	20
1900	12:00	30	20	10	15	15	15	15	20
2000	1:00	30	20	10	15	15	10	15	20
2100	2:00	20	20	10	15	15	10	15	20
2200	3:00	20	20	15*	15	15	10	15	20
2300	4:00	20	20	15	10	15	10	10	20

MID USA

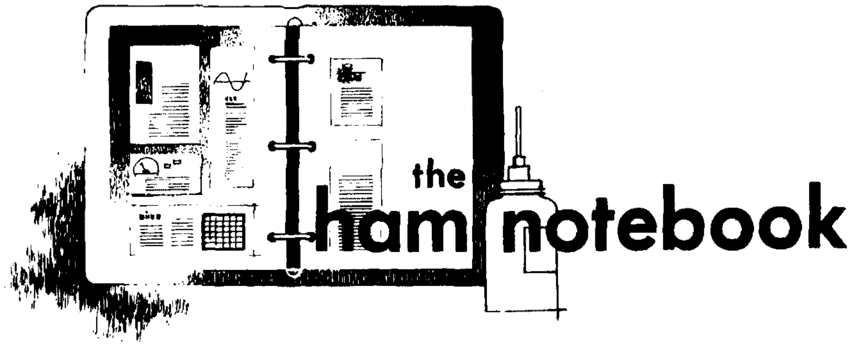
MDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN	CDT
6:00	30	20	15	15	15	10	10	20	7:00
7:00	30	20	15	15	15	10	10	20	8:00
8:00	30	20	20	20	15	15	10	20	9:00
9:00	30	20	20	20	20	15	10	20	10:00
10:00	30	30	20	20	20	15	10	20	11:00
11:00	30	30	20	20	20	15	15	20	12:00
12:00	30	30	20	20	20	20	15	20	1:00
1:00	*30*	*30*	20	*20*	20	20	15	20	2:00
2:00	*30*	*30*	*20*	*20*	*20*	20	20	20	3:00
3:00	*30*	*30*	*20*	*20*	*30*	20	*20*	*30*	4:00
4:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	*30*	5:00
5:00	*20*	*30*	*20*	*15*	*30*	*20*	*20*	*30*	6:00
6:00	*20*	30	20	*15*	*30*	*20*	*20*	*30*	7:00
7:00	*20*	20	15	15	*20*	*20*	*20*	*30*	8:00
8:00	20	20	15	15	*20*	*20*	*20*	*30*	9:00
9:00	20	20	15	10	*20*	*20*	*20*	*30*	10:00
10:00	20	20	15	10	*20*	*20*	20	*30*	11:00
11:00	20	20	15	10	20	15	20	30	12:00
12:00	20	20	10	10	20	15	15	20	1:00
1:00	20	20	10	10	15	15	15	20	2:00
2:00	20	20	10	10	15	10	15	20	3:00
3:00	20	20	10	10	15	10	15	20	4:00
4:00	20	20	15	10	15	10	15	20	5:00
5:00	20	20	15	15	15	10	10	20	6:00

EASTERN USA

EDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE CARIBBEAN S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
8:00	30	20	15	15	15	10	10	20
9:00	30	20	15	15	15	10	10	20
10:00	30	20	20	20	15	15	10	20
11:00	30	20	20	20	20	15	10	20
12:00	30	30	20	20	20	15	15	20
1:00	30	30	20	20	20	15	15	20
2:00	*30*	30	20	20	20	20	15	20
3:00	*30*	*30*	20	*20*	20	20	20	*20*
4:00	*30*	*30*	*20*	*20*	*20*	20	20	*30*
5:00	*20*	*30*	*20*	*20*	*30*	20	*20*	*30*
6:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	*30*
7:00	*20*	*30*	*20*	*15*	*30*	*20*	*20*	*30*
8:00	*20*	30	15	*15*	*30*	*20*	*20*	*30*
9:00	*20*	20	15	15	*20*	*20*	*20*	*30*
10:00	20	20	15	15	*20*	*20*	*20*	*30*
11:00	20	20	15	10	*20*	*20*	*20*	*30*
12:00	20	20	15	10	*20*	15	20	*30*
1:00	20	20	10	10	20	15	20	30
2:00	20	20	10	10	20	15	15	20
3:00	20	20	10	10	20*	15	15	20
4:00	20	20	10	10	15	10	15	20
5:00	20	20	10	10	15	10	15	20
6:00	20	20	15	10	15	10	15	20
7:00	30	20	15	15	15	10	10	20

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during "normal" hours.

*Look at next higher band for possible openings.



RTTY oscilloscope input using line transformers

Here's how to use line transformers ("ouncers") for inputting an RTTY signal to a monitor scope without using any active stages, yet still provide enough gain to produce a picture one and a half inches high.

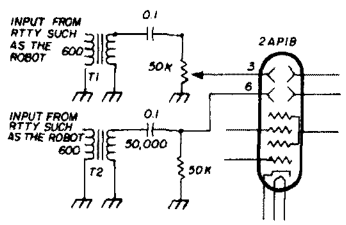


fig. 1. Small line transformers called "ouncers" provide sufficient amplification of RTTY signal to monitor scope input.

The lack of inexpensive monitor scopes for RTTY makes it worthwhile to modify some of the older one- and two-inch oscilloscopes. (Check the surplus market for bargains.) This circuit (**fig. 1**) was applied to an old Millen AM monitor scope. (Circuits for oscilloscopes can be found in the handbooks.[1])

reference

1. *ARRL Handbook*, American Radio Relay League, Newington, Connecticut, 1964, page 544.

Ed Marriner, W6XM

prerecorded messages help the hearing impaired

It's easy to devise a system that enables hearing-impaired persons to communicate with fire, police, ambulance, and other emergency services having TTY or TTY-type-equipment.

Using the TTY machine of the service to be addressed, and a portable cassette tape recorder, record the name, address, phone number, and nature of the emergency to be communicated. Record this data *at least twice* to ensure that all essential information will be transferred, and mark the tape and its container with the name of the emergency described: fire, burglary or assault, or medical emergency, and the phone number of the appropriate service. (If the individual has a particular medical condition, it might be a good idea to prepare an additional tape naming that condition, so that the service can be prepared to respond appropriately in the event of an emergency requiring specialized care.)

This is how it's done:

1. Enter the necessary data into the TTY machine at the headquarters of the emergency service.

2. Set your cassette recorder on "record" and dial the number of the telephone to which data from the TTY will be transmitted. (Do not use the telephone used on the TTY machine.)

3. Hold the microphone near the earphone of the telephone, or attach an inexpensive suction-cup pickup. Record the data from the TTY machine.

4. After recording, send the data back to the TTY machine. Be sure to verify successful transmission.

In an emergency, all the hearing-impaired person needs to do is dial the number, hold the tape recorder to the mouthpiece of the phone and push the "PLAY" button as soon as the call is answered. (To confirm that the phone has been answered, the individual places a fingertip on the diaphragm of the mouthpiece and feels the vibrations of the rings.)

In areas in which TTY facilities are not yet available, the same system can be applied, using prerecorded vocal messages instead of TTY transmissions.

J.W. Dates, W2QLI

modified Bobtail

A modification of the standard Bobtail curtain shown in **fig. 1A** provides good performance on four bands (75,

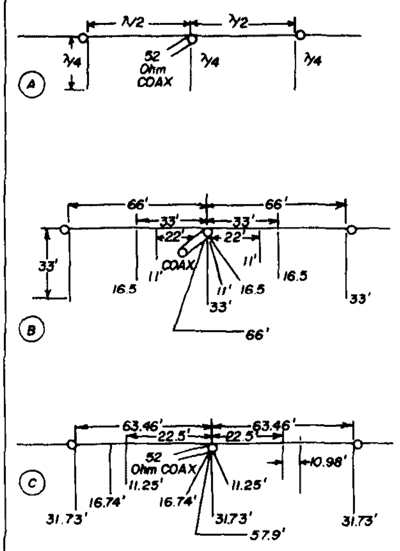


fig. 1. Modified Bobtail curtain.

40, 20, and 15). Center fed with coax, it uses additional lengths of wire placed as shown in **fig. 1B**. The center 66-foot leg can be folded as required if the antenna is lower than 66 feet.

A Navy MARS version of this same antenna is shown in **fig. 1C**. The design — for 4.04, 7.375, 13.975, and 20.8 MHz — requires a height of at least 40 feet above ground. The 57.9-foot radiator for 4.040 MHz, which must be folded at low heights, will require considerable adjustment to resonate on the desired frequency, with the folded portion supported just a few feet above ground level.

The advantage of this experimental antenna is a power gain much higher than a simple dipole on all frequencies above 75 meters, on which the 4.040 MHz radiator functions as a simple upside-down quarter-wave vertical. The center radiators must be kept separated to avoid excessive interaction. The center radiator is fed with 52-ohm coax, preferably through an antenna tuner.

Cliff Francis, W0MBP

fastening Trigon reflectors to VHF antennas

My EME array for two meters uses the method shown below to fasten the Trigon reflectors. It might be well to cut a small V in the rear of the main boom in areas subject to extreme winds. (This was not done on my antenna.) There is no indication of loosening after several windstorms.

The slots through which the hose clamp passes were cut with a saber saw; an ordinary hacksaw blade isn't quite thick enough to provide a slot wide enough to permit easy passage of the hose clamp.

George N. Chaney, W5JTL

VIC-20 printer

It's easy to build an inexpensive printer for the VIC-20 using an ASR-33 teletype machine and the interface illustrated in **figs. 1A** and **1B**.

The printer, which produces typewriter-quality text, won't do everything that an expensive printer will do, but it will allow data listing and, in general, enhance your ability to communicate with your VIC-20. ASR-33's can be found for as little as $50 to $75; other

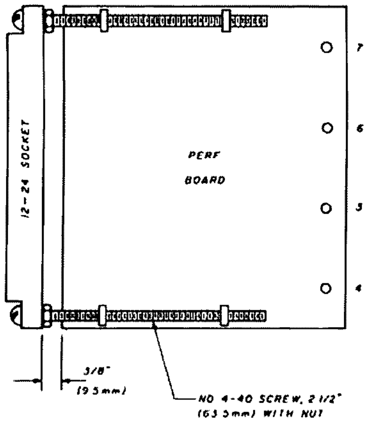
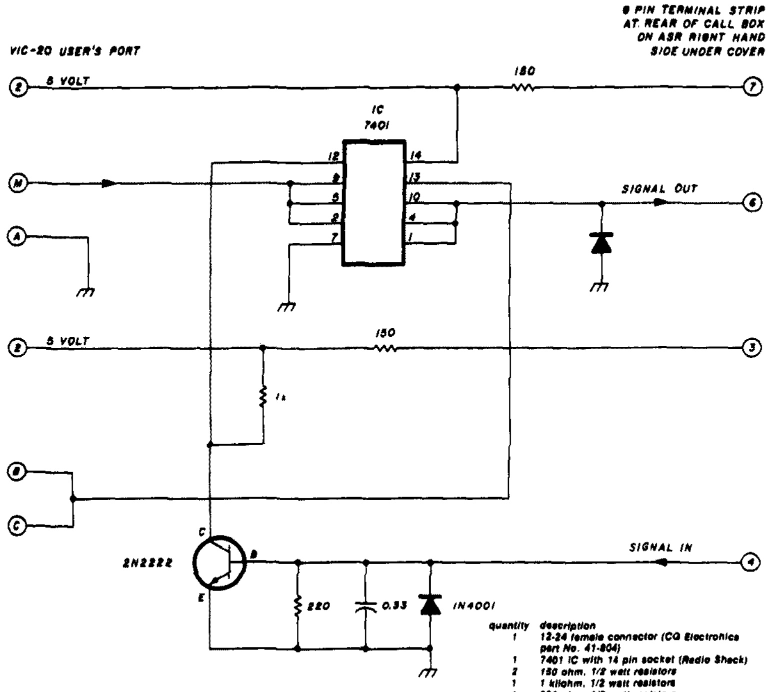


fig. 1A. Interface circuit uses multi-pin edge connector to match VIC-20 requirements.

materials can be found in your junkbox or acquired at little cost. The cost of the entire project should not exceed $100.00.

The interface is inserted into the user's port of the VIC-20, and joined to the ASR-33 by means of a four-wire cable. A simple program (**fig. 2**) provides instructions to the VIC-20.



quantity	description
1	12-24 female connector (CQ Electronics part No. 41-804)
1	7401 IC with 14 pin socket (Radio Shack)
2	150 ohm, 1/2 watt resistors
1	1 kilohm, 1/2 watt resistors
1	220 ohm, 1/2 watt resistors
1	0.33 μF capacitor
2	1N4001 diode
1	2N2222 transistor
1	piece of perforated board 2 × 2-1/2 inches 0.100 × 0.100 hole spacing
2	4-40 × 2-1/2 inch screws with nuts

fig. 1B. Interface circuit that ties a VIC-20 to an ASR-33.

```
10 REM ASR 33 TTY
15REM FILE#>128 FOR CR WITH LF
20REM 163=2 STOP,7 ASCII,110 BAUD
30REM 224=SPACE PARITY,FULL DUPLEX
100OPEN 129,2,3,CHR$(163)+CHR$(224)
110GET#129,A$
200REM MAIN LOOP
220IF B$<>"" THEN IF B$=CHR$(13)THEN
PRINT#129,B$;CHR$(10);CHR$(0);CHR$(
0);:GOTO230
225IF B$<>""THEN PRINT#129,B$;
230GET#129,C$:IF C$<>""THEN PRINT #129,C$;
240PRINT B$;C$;
250SR=ST:IFSR=0THEN220
300REM ERRORS
310PRINT "ERROR";
320IF SR AND 1 THEN PRINT"PARITY"
330IF SR AND 2 THEN PRINT"FRAME"
340IF SR AND 4 THEN PRINT"RCVR BUF FULL"
350IF SR AND 8 THEN PRINT"BREAK"
360IF(PEEK(37151)AND64)=1 THEN 360
370CLOSE 129:END
```

fig. 2. VIC-20/ASR-33 program listing.

J.W. Dates, W2QLI
ham radio

Adjust element spacings, lengths
to improve gain, F/B,
pattern, and bandwidth

applied Yagi antenna design part 5: additional optimization techniques

In the first four parts of this series, specific Yagi antenna designs were optimized for each of four VHF/UHF bands.[1-4] At a designated frequency in each band's weak signal area, computer iterations provided calculated maximas for forward gain and F/B. Three well-known Yagi antenna design approaches formed the bases for these iterations, with comparisons being presented in reference to a standardized fourth approach, the NBS Yagis.[5] For the same Yagi design approach at the same design frequency, some of the more significant findings from these iterations and *comparisons* include the following:

- Parasitic element lengths for maximum calculated forward gain and F/B are different, sometimes significantly so.
- Increasing the tapering of a given design approach requires a longer first director.
- Increases in tapering initially, but not always, result in some increase in calculated forward gain.
- Increased tapering almost always results in some increase in calculated F/B.
- Carefully selected arrangements of unequal director spacing result in more gain for a given boom length, as compared to carefully selected arrangements of equal director spacing, sometimes with fewer directors as well.

Before proceeding further with a general discussion of gain or F/B optimizing techniques, it should be useful to consider why a Yagi antenna produces peaks and/or nulls in both desired and undesired directions.

pattern generation

The Yagi antenna serves to form a traveling wave whose shaping is a function of the amplitude and phasing of the currents in the parasitic elements. These currents result from driving one or more elements with an RF voltage at a given design frequency. Phase relationships are determined by the self impedance of each element and the mutual impedances among the elements. Self impedance is a function of an element's length and diameter, and mutual impedance is a function of the spacing between elements. A Yagi antenna is essentially a system of dipole elements whose resulting radiation patterns are combined into a single pattern. This combining process can be represented in terms of vector quantities that are based on the current flowing in each element. As it is dipole patterns that are being combined, there is usually some resemblance between the resulting Yagi pattern and a single dipole pattern. Some of these resemblances are a peak at zero degrees, a sharp null at 90 degrees, and a lesser peak at 180 degrees. As Yagi patterns are generally symmetrical, a null also occurs at 270 degrees. The patterns in references 1-4 are typical examples.

Forward gain is a measurement of the sharpness of the definition of the main lobe (zero degree lobe). F/B

By **Stanley Jaffin, WB3BGU,** 800 Stonington Road, Silver Spring, Maryland 20902

table 1. Design parameters for a six-element Yagi whose gain is spacing optimized from 11.21 dBi to 12.87 dBi, with a 1.5 dB increase in F/B.

element	element length (λ)	initial spacing from previous element (λ)	optimized spacing from previous element (λ)
reflector	0.51	0.000	0.000
driven	0.50	0.250	0.250
director 1	0.43	0.310	0.336
director 2	0.43	0.310	0.398
director 3	0.43	0.310	0.310
director 4	0.43	0.310	0.407

table 2. Design parameters for a six-element Yagi whose gain is spacing optimized from 10.92 dBi to 12.89 dBi, with a 1.4 dB increase in F/B.

element	element length (λ)	initial spacing from previous element (λ)	optimized spacing from previous element (λ)
reflector	0.51	0.000	0.000
driven	0.50	0.280	0.250
director 1	0.43	0.310	0.352
director 2	0.43	0.310	0.355
director 3	0.43	0.310	0.354
director 4	0.43	0.310	0.373

table 3. Design parameters for a ten-element Yagi whose gain is spacing optimized from 13.07 dBi to 14.25 dBi.

element	element length (λ)	initial spacing from previous element (λ)	optimized spacing from previous element (λ)
reflector	0.51	0.000	0.000
driven	0.50	0.250	0.250
director 1	0.43	0.310	0.319
director 2	0.43	0.310	0.357
director 3	0.43	0.310	0.326
director 4	0.43	0.310	0.400
director 5	0.43	0.310	0.343
director 6	0.43	0.310	0.320
director 7	0.43	0.310	0.355
director 8	0.43	0.310	0.397

Note: While the text in the original source gives the director spacing as 0.310 wavelengths, a table provided in that article gives a dimension of 0.330 wavelengths for director spacing. It is apparent, however, that this Yagi was meant to be an extension of the prior Yagis, with the extra four directors serving merely to show how the model worked with longer Yagis.

is a ratio of the amplitude of this main lobe and the amplitude of the 180 degree lobe. F/B is more critically affected by even slight changes in element current than is forward gain. Variances of many decibels of F/B are often accompanied by forward gain changes that are fractions of a decibel. Overall pattern structure, however, is also materially affected. This explains why forward gain and F/B, while the most popular and perhaps the most significant measures of Yagi performance, do not always accurately reflect a single Yagi's performance or the comparative performances of two or more Yagis.

With three variables to determine current amplitudes and phases in the parasitic elements, it is possible to obtain virtually the same gain or F/B figures with different combinations of these variables. This explains why Yagis with what appear to be measurably different design approaches can have almost the same gain, F/B, or other measures of performance. It is also worth noting that element diameter is usually chosen for reasons of mechanical stability, and is therefore not iterated with the other variables.

Gain or F/B can be mathematically optimized by iterating a single variable, parasitic element length.[1-4] The same results can be calculated by holding parasitic element length constant and iterating parasitic element spacing. For a system of given element lengths, optimum spacings could be found. This could be either in terms of a given number of elements or a given boom length. Likewise, optimal performance values can be calculated for different element lengths when either of these two spacing parameters are held constant. When a given Yagi is described in the Amateur literature as "optimal," it is necessary to ask *what* has been optimized — gain or F/B? It also follows that whenever any of these parameters are changed, "optimal" must now be re-optimized. For example, a seven-element Yagi made by adding an element to an already "optimal" six-element Yagi is not an "optimal" seven element Yagi; the spacings of the other elements (and perhaps their lengths) need to be re-optimized. When two directors were added to the K2RIW Yagi, optimal director length (for gain) dropped from 11.75 to 11.50 inches.[3] Failure to do so would have cost 0.737 dB of gain and 8.784 dB of F/B. It is also necessary to determine if the new boom length might be better served with five (or eight) elements, and if the resulting F/B is available across the entire weak signal area or is a function of single frequency vectorial cancellation.

It would seem logical that if all of these variables could be optimized at the same time, a Yagi antenna with truly phenomenal performance parameters might be designed. The Lawson model used in references 1-4 can easily be adapted to perform these calculations. One series of articles in the professional literature reporting the results of using another model for this same purpose is summarized below.

continuous Yagi antenna performance parameter optimization

A series of computer programs for this purpose has been prepared and apparently successfully executed. An existing Yagi design approach is described in the programs in terms of element spacing, length, and diameter, as well as operating frequency. The program

further iterates the design parameters to optimize the Yagi's gain. Cheng and Chen[6] provide a highly mathematical description of a Yagi element spacing optimization procedure, and give three examples of its use. In the first example (summarized in **table 1**) a six-element Yagi with an initial gain of 11.21 dBi is element space optimized to 12.87 dBi. In the second example (summarized in **table 2**) another six-element Yagi with an initial gain of 10.92 dBi is similarly optimized to 12.89 dBi. In the third example (summarized in **table 3**) a 10-element Yagi with an initial gain of 13.07 dBi is similarly optimized to 14.25 dBi.

For element length optimization, Chen and Cheng[7] provide a description of the process and two examples. In the first example, summarized in **table 4**, a six-element Yagi with an initial gain of 10.89 dBi is element length optimized to 12.15 dBi. In the second example, summarized in **table 5**, this same six-element Yagi is first space optimized to 12.83 dBi, and is then length optimized to 13.40 dBi. (Note: for the Yagis described in **tables 1-5**, element diameter is 0.006738 wavelengths and the booms are non-conductive. F/B calculations were not given, but this figure is derived from the plots provided in both articles.)

A combination of length and spacing optimization or a series of either of these individual optimizations can be continued. The rapid convergences described in both articles indicated that very little additional gain would be realized. If gain figures for various boom lengths are extrapolated from the NBS findings in reference 5, they can be used to arrive at very favorable comparisons with the Chen-Cheng results. Yagis whose gain figures were initially low for comparable NBS Yagis of equal boom lengths, were optimized to gain figures equal to or in excess of these same NBS figures. The accuracy of these results depends on a careful validation of the Chen-Cheng model against all of the NBS Yagis, or against some other equally validated reference. Unfortunately, this could not have been done because the NBS data was

table 4. Design parameters for a six-element Yagi whose gain is length optimized from 10.89 dBi to 12.15 dBi, with a 1.6 dB increase in F/B.

element	spacing from previous element (λ)	initial element length (λ)	optimized element length (λ)
reflector	0.000	0.510	0.472
driven	0.250	0.490	0.456
director 1	0.310	0.430	0.439
director 2	0.310	0.430	0.444
director 3	0.310	0.430	0.432
director 4	0.310	0.430	0.404

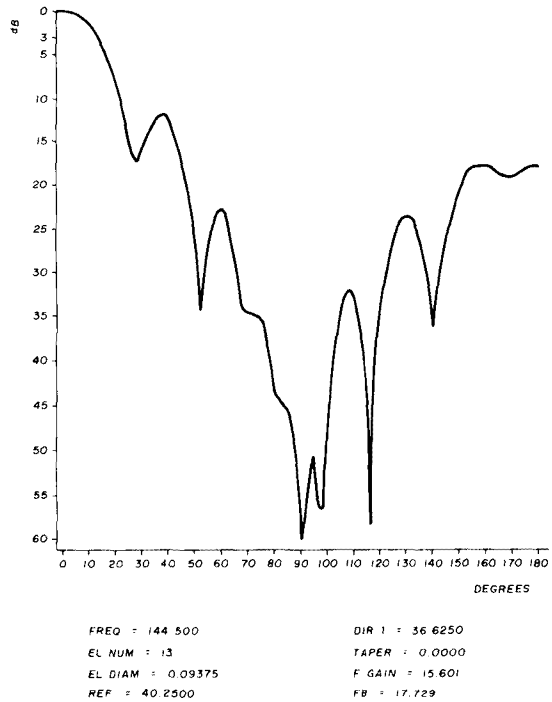


fig. 1. E-plane plot of the zero-taper, gain optimized Kmosko-Johnson Yagi antenna.

table 5. Design parameters for a six-element Yagi whose initial gain of 10.89 dBi is space optimized to 12.83 dBi and then length optimized to 13.40 dBi, with a 0.4 dB increase in F/B.

element	initial design parameters: spacing from previous element (λ)	initial design parameters: element length (λ)	optimized design parameters: spacing from previous element (λ)	optimized design parameters: element length (λ)
reflector	0.000	0.510	0.000	0.476
driven	0.250	0.490	0.250	0.452
director 1	0.310	0.430	0.289	0.436
director 2	0.310	0.430	0.406	0.430
director 3	0.310	0.430	0.323	0.434
director 4	0.310	0.430	0.422	0.430

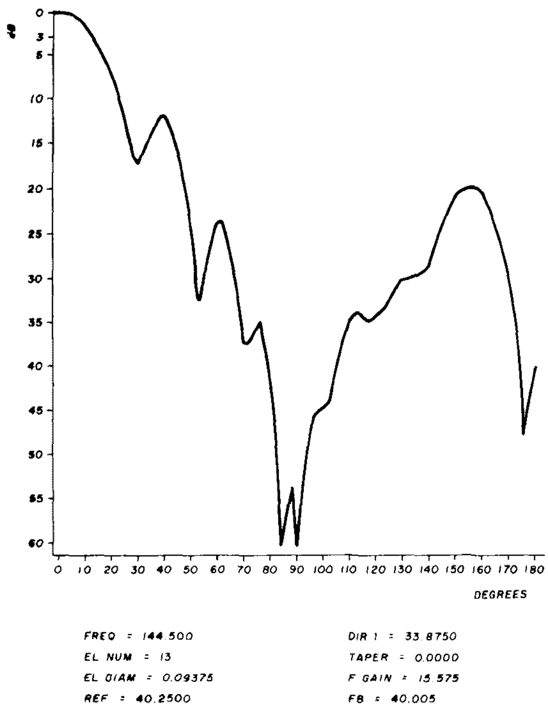


fig. 2. E-plane plot of the F/B maximized version of the zero-taper, gain optimized Kmosko-Johnson Yagi antenna.

not published until three years after Chen and Cheng's first article. Comparisons with the Lawson model on the Yagis contained in the two articles would not be a conclusive test. A Yagi model needs to be validated against a wide range of parasitic element numbers and spacings, not a select few that might fall through some mathematical "cracks."

Articles in the Amateur literature also describe techniques that have been shown to increase Yagi gain or F/B. These techniques involve adding another element or altering the length of an existing element. The following two examples given are from the sources cited in references 1-4.

F/B optimization for the Kmosko-Johnson Yagi

Reference 1 described an extensive analysis of the Kmosko-Johnson design approach for long Yagis. These computer iterations showed how increased director tapering resulting in increased gain and significantly increased F/B. Kmosko and Johnson's original article made mention of a rather unique method for increasing F/B,[8] in which the last director is made shorter than its tapering schedule would normally require, resulting in a higher F/B ratio. While Kmosko and Johnson mentioned a very slight decrease in this director's length, computer iterations were made over a wide range of such decreases.

The last director of each of four gain optimized Kmosko-Johnson Yagis was continually decremented by 0.0625 inch. These Yagis had director tapering schedules of 0.000, 0.0625, 0.125, and 0.1875 inch. The most dramatic results were obtained for the zero taper Yagi, giving further credence to these designers' belief (and the findings in reference 1) that their Yagi performed better with at least some degree of director tapering. **Table 6** presents the results for the zero taper gain optimized Yagi, with the initial gain optimized performance parameters followed by those measured at 0.25-inch decrements (every fourth 0.0625 decrement). **Table 7** and **8** present comparisons between the original Yagi and the Yagi optimized for F/B by this process. **Figs. 1** and **2** present the E-plane plots for these respective Yagis.

The newly optimized F/B derives from single frequency vectorial cancellation. There is nearly a 16 dB drop across the 144-MHz weak signal area. The gain remains nearly constant, but at a level below that of the original Yagi. The newly optimized Yagi has a different lobe structure in its second quadrant (90-180 degrees). The overall reduction of signal pick-up is greater than that of the original optimized Yagi, and from 160-180 degrees, the nearly 23 dB increase in F/B becomes readily apparent. As is usually the case with antennas, the selection of either of these Yagis is a matter of the station operator's personal preference. Reference 1 provides the potential user of Kmosko-Johnson Yagis with additional alternatives.

F/B optimization for the Tilton/Greenblum Yagi

References 2-4 described many Yagis based on this proven design. In Greenblum's first article, he mentions a method for increasing the F/B ratios of the Yagis described in his design tables.[9] This involves adding another director, but only after determining its position by moving it down the boom to find the F/B maxima. The Greenblum design is based on finding the gain maxima, as Greenblum was obviously concerned about F/B and the overall pattern.

The 0.000 taper gain optimized Yagi from reference 2 was selected for the computer iterations used to illustrate this technique. With tapered directors there would be the additional problem of controlling for a second variable, the taper of the new director. All iterations were made at the Yagi's design frequency of 220.5 MHz.

In order to avoid having this extra director coincide with any of the existing directors, an initial spacing from the reflector of 0.15 wavelengths was selected. Increments were in steps of 0.01 wavelengths, with the last positioning of this director being 3.20

table 6. The effects of decreasing the length of the last director on the performance parameters of the gain optimized zero taper Kmosko-Johnson Yagi.

length of last director (inches)	gain (dBi)	F/B (dB)
36.625	15.601	17.729
36.375	15.604	19.090
36.125	15.605	20.525
35.875	15.604	22.054
35.625	15.602	23.702
35.375	15.599	25.509
35.125	15.595	27.531
34.875	15.592	29.851
34.625	15.587	32.588
34.375	15.583	35.840
34.125	15.579	39.164
33.875	15.575	40.005
33.625	15.571	37.681
33.375	15.567	35.048
33.125	15.563	32.923
32.875	15.559	31.249
32.625	15.555	29.905
32.375	15.552	28.800
32.125	15.548	27.873

table 7. Frequency response parameters for the initial zero taper gain optimized Kmosko-Johnson Yagi.

frequency	gain (dBi)	F/B (dB)
142.5	15.259	11.162
143.0	15.344	12.022
143.5	15.440	13.266
144.0	15.537	15.126
144.5	15.601	17.729
145.0	15.524	18.591
145.5	15.082	13.846
146.0	13.920	8.321
146.5	11.736	3.390

table 8. Frequency response parameters for the initial zero taper gain optimized Kmosko-Johnson Yagi after maximizing F/B by reducing the length of the last director.

frequency	gain (dBi)	F/B (dB)
142.5	15.301	14.600
143.0	15.404	16.310
143.5	15.500	19.015
144.0	15.570	24.302
144.5	15.575	40.005
145.0	15.434	20.942
145.5	15.013	13.968
146.0	14.137	9.146
146.5	12.653	5.125

wavelengths from the reflector. A five-page listing, detailing many undulating cycles of both gain and F/B results, was produced. **Table 9** presents comparisons among the performance calculations for the original Yagi, a new gain optimized Yagi, and each of the Yagis at calculated F/B maximas. **Tables 10, 11,** and **12** present the frequency response characteristics for the original Yagi, the new gain optimized Yagi, and the F/B optimized Yagi resulting from this process. **Figs. 3, 4,** and **5** present these antenna's respective E-plane plots.

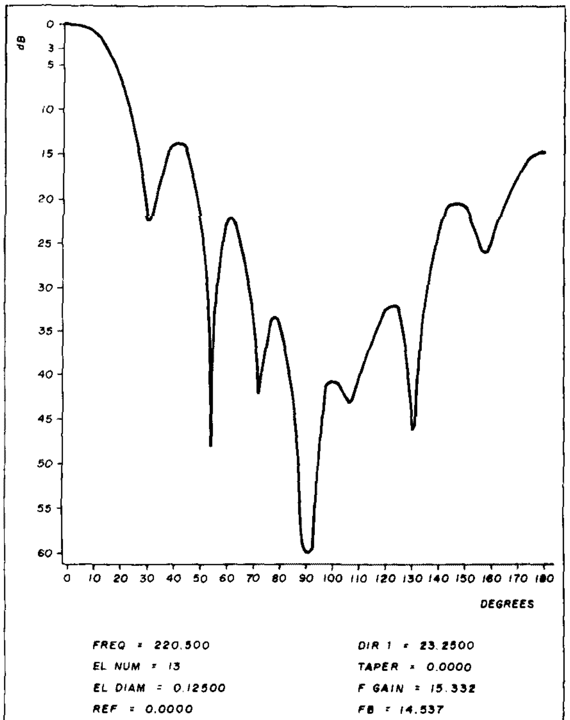


fig. 3. E-plane plot of the zero-taper, gain optimized Tilton-Greenblum Yagi antenna.

table 9. Performance parameter comparisons among various gain optimized zero taper Tilton/Greenblum Yagis as a function of placement of an additional director.

spacing of extra element from reflector (λ)	gain (dBi)	F/B (dB)	comments
---	15.332	14.537	original Yagi
0.28	15.476	16.030	local F/B maxima
0.35	15.576	13.614	new gain maxima
0.69	14.482	21.392	local F/B maxima
1.17	14.551	23.888	local F/B maxima
1.67	14.878	19.850	local F/B maxima
2.12	14.872	21.325	local F/B maxima
2.61	14.611	29.760	global F/B maxima
3.07	15.116	19.931	local F/B maxima

With the exception of the amplitude of some of the minor lobes, there are no real differences between the *original* and the new *gain optimized* Yagis. The F/B optimized Yagi has the minor lobes with the greatest

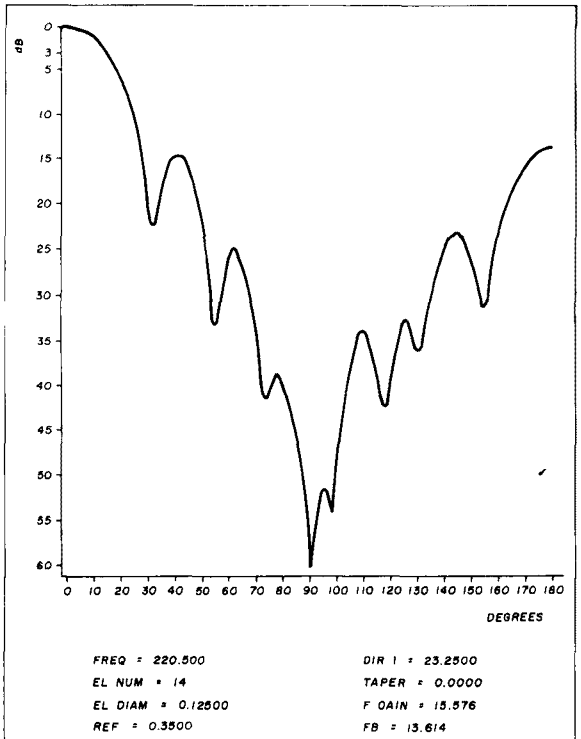


fig. 4. E-plane plot of the new zero-taper, gain optimized Tilton-Greenblum Yagi antenna.

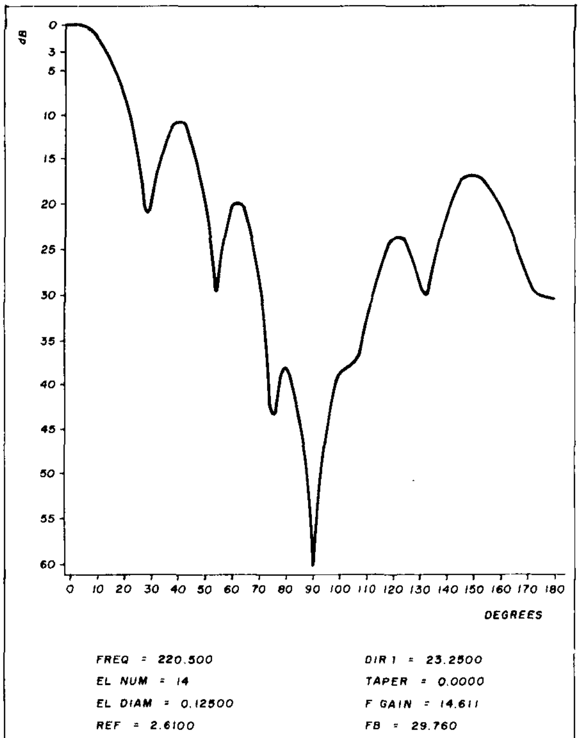


fig. 5. E-plane plot of the F/B maximized verison of the zero-taper, gain optimized Tilton-Greenblum Yagi antenna.

table 10. Frequency response parameters for the initial zero taper gain optimized Tilton/Greenblum Yagi.

frequency	gain (dBi)	F/B (dB)
216.5	14.737	16.075
217.5	15.007	16.472
218.5	15.194	16.169
219.5	15.301	15.425
220.5	15.332	14.537
221.5	15.295	13.695
222.5	15.201	13.003
223.5	15.065	12.515
224.5	14.903	12.266

amplitude, particularly from 100-160 degrees. It is only from 165-180 degrees that the reduced signal pick-up associated with a higher F/B is apparent. All three Yagis display an F/B that increases at frequencies higher than the design frequency, a characteristic of a Yagi inherently optimized for something other than F/B. Specially placing an element to optimize F/B has not changed this Yagi's basic performance characteristics. While individual preferences are generally an important factor in Yagi selection, the 220 MHz operator desiring a high F/B may be better served with those Yagis presented in reference 2.

concluding comments

Throughout this series I have emphasized the use of the digital computer as an antenna design tool capable of providing the VHF/UHF Radio Amateur with a wealth of accurate information in a relatively short time. Gone are the days of tedious and seemingly endless iterations of element lengths, spacings, and resulting pattern measurements on antenna test ranges, all with the inherent possiblity for significant human error. With the computer, several "lifetimes" of Yagi design iterations can be performed accurately and painlessly by using a model that starts with specific designs. More importantly, the VHF/UHF operator can estimate the expected performance of any design more closely than has been possible in the past.

This series has shown how to identify Yagis whose overall performance parameters — a well defined main lobe, reduced side lobes, and a reasonable F/B ratio — are most desirable in the VHF/UHF station. Once again, Reisert has restated and illustrated the importance of emphasing these parameters.[11] However, the operator with special needs, or with the age-old urge to tinker with antennas, can now do so with relative ease.

To further aid the knowledgeable antenna experimenter, the final article in this six-part series will pre-

table 11. Frequency response parameters for the new gain optimized zero taper Tilton/Greenblum Yagi with an additional director 0.35 wavelengths from the reflector.

frequency	gain (dBi)	F/B (dB)
216.5	14.850	14.071
217.5	15.171	14.322
218.5	15.391	14.215
219.5	15.521	13.920
220.5	15.576	13.614
221.5	15.573	13.441
222.5	15.528	13.529
223.5	15.451	14.025
224.5	15.330	15.149

table 12. Frequency response parameters for the new F/B maxima on a gain optimized zero taper Tilton/Greenblum Yagi with an additional director 2.61 wavelengths from the reflector.

frequency	gain (dBi)	F/B (dB)
216.5	14.322	16.928
217.5	14.504	19.378
218.5	14.613	22.262
219.5	14.650	25.709
220.5	14.611	29.760
221.5	14.491	33.654
222.5	14.279	33.505
223.5	13.958	28.425
224.5	13.493	22.647

sent a detailed explanation of the FORTRAN program that enables the mathematical model to be iterated on a digital computer. An NBS Yagi will be used to illustrate the program's logic, and a copy of the FORTRAN program will be made available at that time.

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ham radio

the *SEED* antenna — a *S*hort, *E*fficient *E*nd-fed *D*ipole

Achieve good efficiency while covering 80/160 meters

In 1978 I needed a **very short antenna** for the 80 and 160 meter bands. Because I was not satisfied with popular designs, I developed a very small antenna which, I hoped, might be better. For the last several years, the result of this effort has been my regular station antenna. I can find nothing like it described in the Amateur literature.

The antenna measures 20 feet (6 meters) long, and its center is 14 feet (4.2 meters) above the ground. Input SWR is less than 1.15:1 from 3.5 to 4.0 MHz (160 meter data is comparable).

If the interest in small antennas is as widespread as it appears to be, and if the numerous popular designs are as inefficient as I believe them to be, then the story of the SEED antenna, whether useful, amusing, or controversial — might be worth telling. My design considerations, measurement techniques, performance data, and evaluation are included: if they were flawed in any way, experts are welcome to set the record straight, but the antenna *does* work . . . and it has advantages I've never seen in any other ham antenna, large or small.

initial design

The first task was to decide what features were considered most important, and what their order or priority should be. They were:

- operation in the 80 and 160-meter bands
- small size
- efficiency
- feed point impedance of 50 ohms
- simplicity of operation

The first two would be mandatory; optimizing the others would be the job at hand.

Since short radiators have low radiation resistance, good efficiency requires even lower loss resistance. The possibilities of a vertical monopole antenna operated against ground were not explored because an adequate radial system is not small and, without one, ground losses are excessive. Similarly, capacitive end loading structures, to be effective, would also be too big. A short dipole, fed at its center, would require loading inductance, which would necessarily have too much resistance to be acceptable.

However, when viewed from its ends, a short dipole exhibits inductive reactance. This can be resonated with capacitance, and the losses in capacitors can be quite small.

The efficiency of such a short antenna could also be enhanced if the radiation resistances were maximized. This could be done by causing more current to flow in a greater part of the length of the radiator. The end-fed design would provide maximum current in the full length of the radiator and maximum radiation resistance for the length available. The antenna would then be a short, efficient, end-fed dipole.

With these thoughts in mind, a 20-foot (6.1 m) piece of 1-inch (25.4 mm) copper pipe was selected for the radiating element corresponding to a length of 0.08 wavelength or 29 degrees at 3.950 MHz, and half that at 1.9 MHz. A tapered transmission line of two 20-foot pieces of 1/2-inch (12.7 mm) copper pipe are used to end feed the driven element while providing closely spaced points for connection of the other parts. This approximates an equilateral triangle, with the narrow end of the transmission line separated by only a few inches. At this point the reactance would still be highly inductive.

Air dielectric tuning capacitors from each side of the transmission line establish resonance. By placing a low

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reactance loading capacitor between them, the two connecting points provide a low impedance, balanced feed point. By proper selection of the loading capacitor value, this becomes a 200 ohm feedpoint, nonreactive when the system is tuned to resonance, and appropriate for a conventional 4:1 balun to match to 50-ohm coaxial cable.

Inherent in this design philosophy is the distinction between the radiator and the transmission line functions. The radiator is a linear elementary dipole in which current is essentially the same at all points along its length, and voltages at its ends are equal and opposite in phase.

The transmission line section consists of two adjacent, straight conductors in which current and voltage are equal and opposite. It is tapered, and therefore its characteristic impedance varies throughout its length. Its center line is perpendicular to the radiator, and all elements lie in the same plane.

If the radiator is mounted vertically, its intrinsic radiation will be vertically polarized and maximum toward the horizon, while the center line of the transmission line will be horizontal, as will the polarization of its radiation.

The inductive reactance present at the open end of the transmission line section is the combination of that at the ends of the radiator and the effects of the line itself. This total was considered the inductive component of the resonant circuit, and no attempt to separate the factors appeared to be necessary.

construction

The antenna assembly is illustrated in **fig. 1**. The pipe was joined using standard soldered plumbing fittings to minimize junction losses. The ends of the transmission line were connected with half inch silver plated braid to rather large feedthrough insulators on the box containing the other parts. A wooden "T" frame supports the pipe and box, and is mounted on a 14-foot (4.27 m) high wooden 4 × 4 inch (100 × 100 mm) pole. The copper pipe weighs about 18 pounds (8.16 kg) and is not self supporting.

A weatherproof box at the end of the transmission line houses the capacitors, balun, and selsyn. The circuit within the box is shown in simplified form in **fig. 2**.

The tuning capacitors, "ganged" by a shaft coupling, are controlled from the operating position by a pair of surplus selsyns connected by a multi-conductor cable. A small reversible, slow speed motor might have been a better choice. The loading capacitors are mounted and connected with copper strap and banana plugs.

The assembly is mounted with the radiating element vertical. The supporting boom is hinged at the top of the pole to allow it to be tilted 90 degrees to bring the control box down to shoulder level for substitution of capacitors during evaluation.

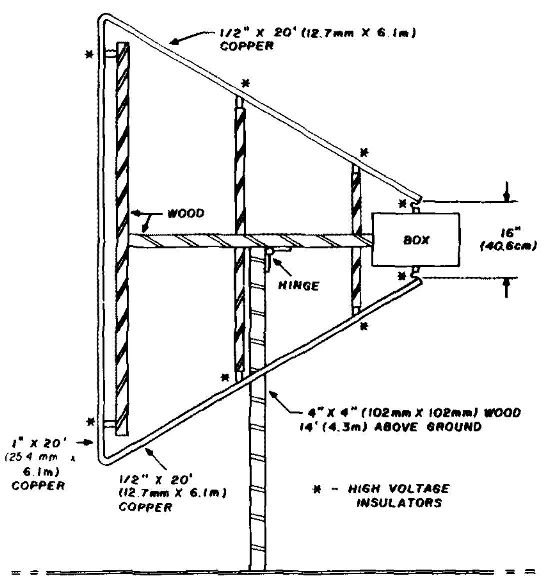

fig. 1. The structure of the SEED antenna and its mounting (not to scale).

After the initial selection, the loading capacitor does not require adjustment to provide a low SWR across a single band. However, a different value is required for each band. Because of the high current to be carried, use of relays or switches is avoided and plug-in units are substituted when changing bands.

circuit description

The basic circuit as originally envisioned is a parallel resonant circuit as shown schematically in **fig. 3A**. The series resonating capacitors, C_{S1} and C_{S2}, and the loading capacitor, C_L, all in series, are across the inductance of the pipe structure, L. There is also a significant distributed capacitance, C_D, across the inductor. This is the capacitance between the sides of the pipe structure plus the stray capacitance of leads to and within the component box. The inductance is 20 μH and the distributed capacitance is 19 pF. The resistance, R, is the sum of the radiation resistance and the loss resistance of the pipe (including joints) and capacitors.

Selection of the series resonating capacitors determines the operating frequency. The value required for the loading capacitor will depend, to some extent, on the physical characteristics of the antenna structure, its mounting and environment, and the adjacent ground. The balun is a standard commercial unit with a cylindrical core and an impedance ratio of 4:1, and is rated for full Amateur power.

The practical circuit now in use is shown in **fig. 3B**. C_{S1} and C_{S2} are, in fact, series or parallel connected assemblies of fixed units, as required by band selec-

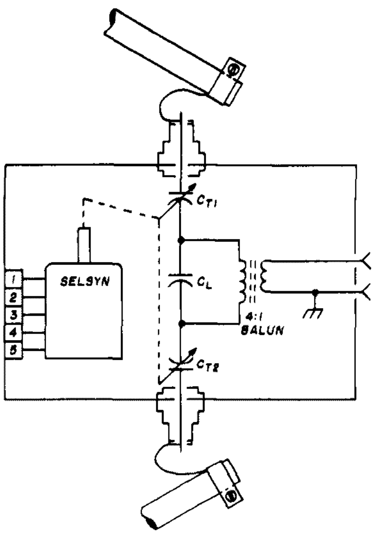

fig. 2. Simplified diagram of the contents of the control box on the boom of the antenna.

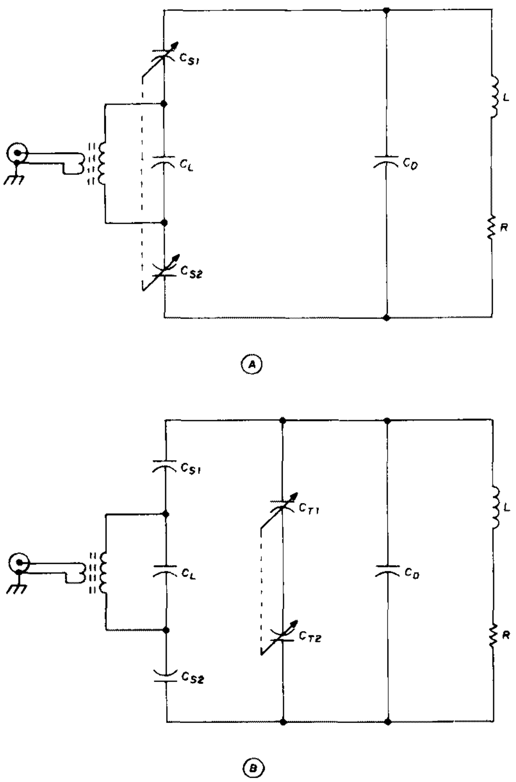

fig. 3. Circuit of the SEED antenna: (*A*) the elementary circuit originally envisioned, and (*B*) the practical circuit in use. (Component designations are explained in text.)

tion and parts availability. In early experiments, each of the variable tuning capacitors was connected in parallel with the related series resonating capacitor. This isolated the distributed capacitance, C_D, and facilitated its measurement. Under these conditions, SWR was 1.5:1 or less from 3.5 to 4.0 MHz. It later developed that connecting the tuning capacitors in series across the entire circuit, as shown, would significantly improve the SWR at the band edges.

Under operating conditions, the total resistance, R, also included the effects of ground and environment. This measured 0.64 ohms at 1.9 MHz and 2.21 ohms at 3.950 MHz.

The tuning capacitors are 10-100 pF, 4500 volt units. When each is paralleled with a fixed series capacitor of 680 pF at C_{S1} and C_{S2}, the circuit tunes from 1.813 to 1.907 MHz. The optimum value of C_L was 7450 pF, and maximum SWR was 1.3:1 in this range. Bandwidth for an SWR of 2:1 without retuning was 3.75 kHz with a loaded Q of about 370. At 200 watts to the antenna, the tuning capacitors each had 2000 volts, RMS, across them and the current in the circuit was 18 amperes, RMS. The benefits of the revised connection of the tuning capacitors had not yet been recognized when 160 meter tests were made.

Using the same tuning capacitors but in the revised circuit, and with fixed 100 pF units at C_{S1} and C_{S2}, the circuit could be tuned from 3.300 to 4.095 MHz. The optimum value of C_L was 1250 pF for an SWR of less than 1.15:1 from 3.5 to 4.0 MHz. The bandwidth for an SWR of 2:1 without retuning was 14 kHz, with a Q of about 225. At 200 watts to the antenna, each tuning capacitor had 2250 volts, RMS, across it, and current in the circuit was 9.5 amperes, RMS.

2-meter model

A 1:36 scale model of the design was made and operated in the 2-meter band in an effort to determine the free space radiating characteristics of the design. Under much less than ideal conditions, scans of 360 degrees of azimuth were made for both horizontally and vertically polarized radiation, with the antenna in three attitudes.

The most informative patterns occurred when the radiator was horizontal and the center line of the feedline section vertical, as shown in **fig. 4A**. This separated the horizontal radiation of the radiator from the vertical radiation of the feedline section, and facilitated consideration of each separately from the other. It emphasized that the maximum signal from the radiator was at right angles to it, whereas that from the feedline section was concentrated in the directions in the plane of the structure and perpendicular to its center line.

With the plane of the assembly horizontal, a plot of the horizontal radiation, as shown in **fig. 4B**, shows essentially a circular pattern, decreasing about 2 dB

off the ends of the radiator. Radiation from the feedline section nearly fills in the nulls at the ends of the radiator. Vertical radiation was not detectable in any direction.

With the radiating element vertical, the vertical radiation pattern, as shown in **fig. 4C** shows lobes in directions in the plane of the antenna which were about 6 dB above the nulls at 90 degrees from them (broadside). There was no measureable horizontal radiation.

orientation

The model tests simulated operation of the SEED antenna in free space. To that information must be added the effects of the proximity of ground. Even though they could not be measured with available facilities, the nature and relative magnitude of the distortions to be expected could be estimated.

Mounted horizontally, the SEED design might be an excellent antenna if it were about 140 feet (42.67 m) above ground. At a height appropriate to its size, its radiation resistance would be reduced, decreasing efficiency. Very little low angle radiation would exist. Ground losses would be severe.

If it were mounted with the plane of the structure vertical and the radiator horizontal, the horizontal radiation would be degraded as described above. The feedline section would produce some vertically polarized radiation.

By mounting the SEED with the radiating element vertical, ground losses might be less and low angle radiation should be improved. Since selection of polarization could not be based on comparison of measurable losses, vertical polarization was chosen to favor lower vertical radiation angles.

initial observations

A unique feature of the SEED design is a feed point

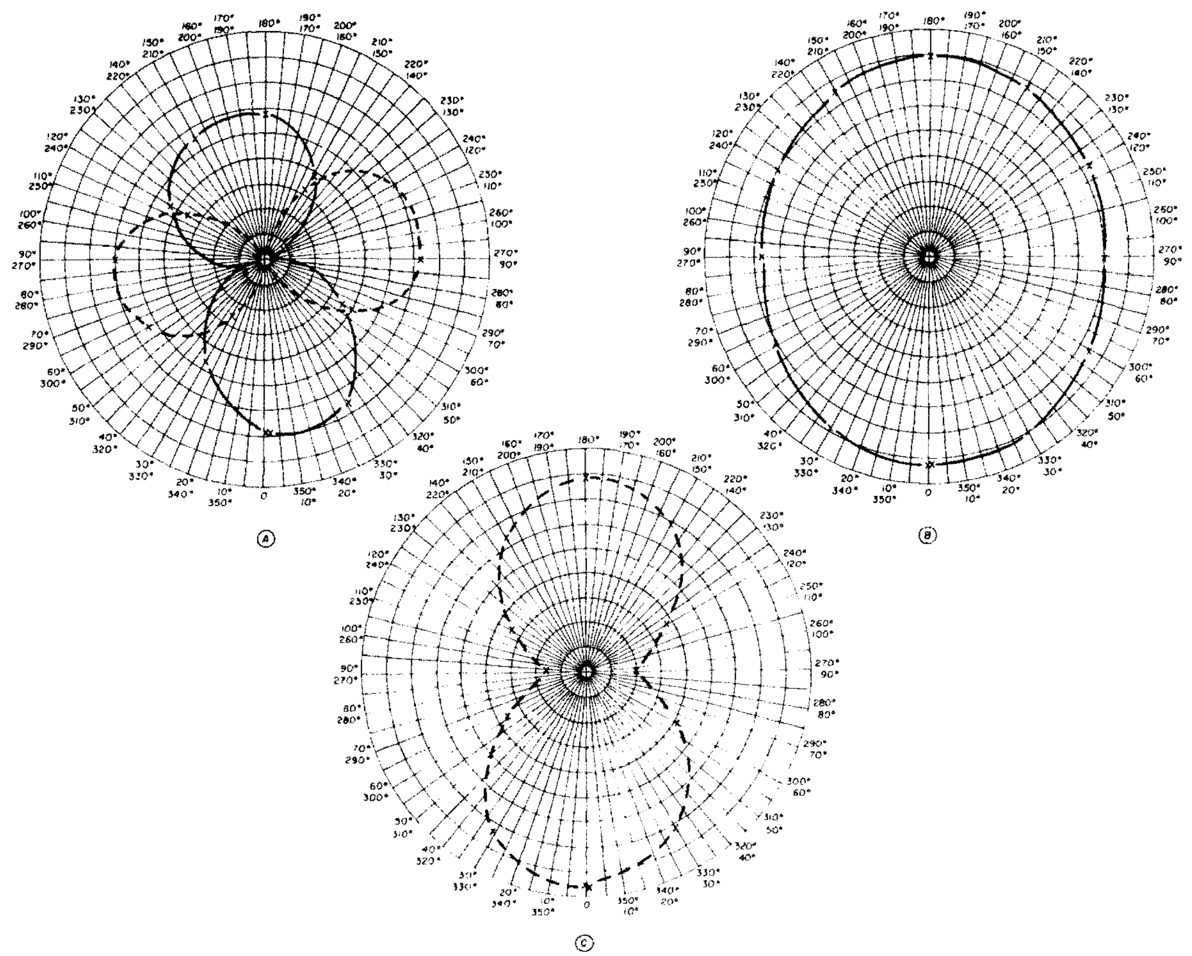

fig. 4. Free space radiation patterns of the 2-meter model SEED: (*A*) with plane of antenna vertical and radiator horizontal, (*B*) with plane of antenna horizontal, and (*C*) with radiating element vertical. Zero degrees in *A* is perpendicular to the plane of the antenna; in *B* and *C*, perpendicular to the radiator in the plane of the antenna, feed point at 180 degrees. Solid line shows horizontally polarized radiation, dashed line, vertical.

impedance of exactly 50 ohms, non-reactive. Other resistive or complex impedances may be obtained if desired. The resistive component is continuously variable by adjusting the value of the loading capacitor, and reactance may be introduced or eliminated by the main tuning control. With the loading capacitor optimized at mid-band, SWR did not exceed 1.15:1 from 3.5 to 4.0 MHz.

Since the loading capacitance is "set-and-forget," only one operating control is needed. A noise bridge or other low power indicator of resistance and reactance at the operating position will show when the antenna is resonant at the desired frequency, and causes no harmful interference. Under power, any device which will show maximum forward or minimum reflected power in the feed line will indicate proper tuning. But accurate tuning is critical to optimum antenna performance as well as feed point impedance. Error in tuning of 8 kHz at 3.950 MHz results in SWR of 2:1 and degrades efficiency, and at 160 meters much more care is necessary. CAUTION: A matching network or "antenna tuner" should not be used with the SEED antenna; neither it nor any controls in the transmitter can compensate for mis-adjustment of this antenna.

If a slow speed, reversible motor is used for remote tuning, a drive shaft speed of 1 RPM is a little too fast for convenience and accuracy, while a slower rate increases the time required for wide frequency excursions.

The very high loaded, operating Q of this circuit, 225 at 3.950 MHz and 370 at 1.900 MHz, probably attenuates harmonics and many other spurious emissions very effectively, but this effect could not be assessed. The resulting high current in the full length of the radiator is the good news. High current and voltage in the other parts of the circuit require special attention. Many hams may not be familiar with antenna parameters of 16,000 volts, peak, or 42 amperes of RF.

No inherent frequency limitations on the SEED design were observed. The 144 MHz model performed well, but both selection and adjustment of low loss, small capacitors were tedious. The total length of the radiator plus both elements of the feedline section should not exceed about 0.4 wavelength at the highest frequency to be used for fundamental operation.

At lower frequencies, through the broadcast band and below, it appears that a structure of this design, but still less than 0.1 wavelength long, would operate well and might have advantages. Elimination of the need for an extensive field of ground radials as an integral part of the circuit may be beneficial in some cases.

operational testing

The test site for the SEED is in a ravine nearly surrounded by ground 130 feet higher. The surface slopes about 8 degrees and is completely covered by trees. There are seven houses within a half wavelength of the antenna. A full length horizontal dipole 35 feet high and a 48-foot vertical are available for comparison. A single knob permits instant selection of any antenna and disabling of the others so they will not act as parasitic radiators. All antennas were matched to accept the same power. Most tests were conducted at frequencies near 3.950 MHz.

A lengthy effort was made to obtain dependable numerical comparative performance data, but results were inconclusive. Subjectively, less formal signal reports and innumerable listening tests over a three-year listening period were encouraging. At distances of less than 100 miles, the consistent superiority of the horizontal dipole confirmed the predominantly low-angle radiation of both the SEED and the vertical. At distances up to about 550 miles, the SEED and the horizontal dipole exceeded each other as conditions varied, while the vertical whip was consistently inferior. The immediate terrain prevents investigation of the probable superiority of the SEED and the vertical at greater distances, where predominantly low-angle radiation is most effective.

The antenna was resonated and matched in the 160 meter band and operated for about three weeks in early April. Power to the antenna was about 160 watts, PEP, on Single Sideband. Most contacts were made between 6 and 10 PM and at distances of 300 to 600 miles. No apology was offered for the size of the antenna and good reports were received. Those who asked and were told that the radiator was 20 feet long expressed surprise and curiosity.

The most frustrating aspect of these experiments was the inability to obtain satisfactory "on the air" performance data. It is hoped that someone with a suitable test site will investigate and report the low angle, long distance capability of the SEED which could not be determined at this location.

measurements

Several years of dredging at surplus outlets and hamfests had provided a supply of nondescript capacitors for this project. It soon became apparent that knowing the capacitance of those in the circuit would be necessary, and accuracy would be important. A Dynascan digital capacitance meter was obtained and used for measurements. A popular noise bridge was found to be inadequate for critical, repeatable, measurements. By modifying a published design, a noise bridge with suitable accuracy and resolution was made and calibrated. A secondary station receiver was dedicated to the project, and a signal generator and frequency counter provided signals of known amplitude and frequency.

One of the useful features of the SEED design is that it is a parallel resonant circuit with easily measurable

components. The series, loading and tuning capacitors can be measured directly. The value of the distributed capacitance and the inductance of the structure, including the connecting leads, can be computed directly from these measurements and the frequency of resonance, as explained in **appendix A**.

With this data, it is possible to determine the series resistance of the antenna in position, as shown in **appendix B**. This is the sum of the radiation resistance, the loss resistance of the components, and the effects of absorption and reflection of adjacent ground and other objects. The loaded Q of the circuit can also be found, and measurement of the bandwidth for an SWR of 2:1 can be confirmed as in **appendix C**.

efficiency

The books say that the radiation resistance at 3.950 MHz of the SEED antenna in free space would be about 5 ohms. Loss resistance in the primary circuit has been determined to be about 0.1 ohm. Under these conditions, efficiency would be $\frac{5}{5 + 0.1} \times 100 = 98$ *percent*. This shows success in obtaining some of the design and construction goals, but usefulness of the figure is limited.

When the real secondary effects of ground and environment loss and reflection are added, the measured total resistance is 2.21 ohms at 3.950 MHz. Facilities were not available to divide this total between loss and radiation resistance, so a real efficiency percentage cannot be determined. However, the portion which represents total losses cannot exceed 2.21 ohms and must be somewhat less because useful radiation has been observed.

This justifies the original premise that even small amounts of loss resistance in the primary circuit could seriously degrade the output signal. As an example, substitution of No. 10 wire for the pipe which was used would increase primary resistance by 1.1 ohms and would just about double the total loss of the operating antenna. In other words, the increased resistance of the wire would be more or less equivalent to the power of the radiated signal.

evaluation

The initial design goals were all satisfactorily met or exceeded. A 20-foot radiator which can be operated without apology on either 160 or 80 meters is certainly small, and the SEED appears to be much more effective in use than would be expected. An SWR of less than 1.15:1 across the entire 3.5-4 MHz band may be one of the best solutions available for modern transceivers with transistor output stages. Elimination of the usual "antenna tuner" in the shack, and a single control to resonate the system, is operating convenience which approaches the ultimate (automatic sensing and resonating of the circuit could be provided).

However, CAB's Law is that "A problem solved is a problem created." At full legal Amateur power, a very short antenna must carry very high current, as do the associated capacitors. In the process of developing that current, high voltages are created. Ham-type handbooks and reference data available here provide very little theoretical or practical information about capacitors under conditions of high voltage, high current, high frequency, and high capacitance values.

The SEED antenna operates at 200 watts on either band with surplus capacitors that cost less than $25. The air variables are currently advertised at about twice that, but all others are obsolete and not now available. The capacitors in SEED Mk IV, a rather different version, are more than is needed, but they will take a full kilowatt without strain. They could be bought new, now, but for well over $1,000.

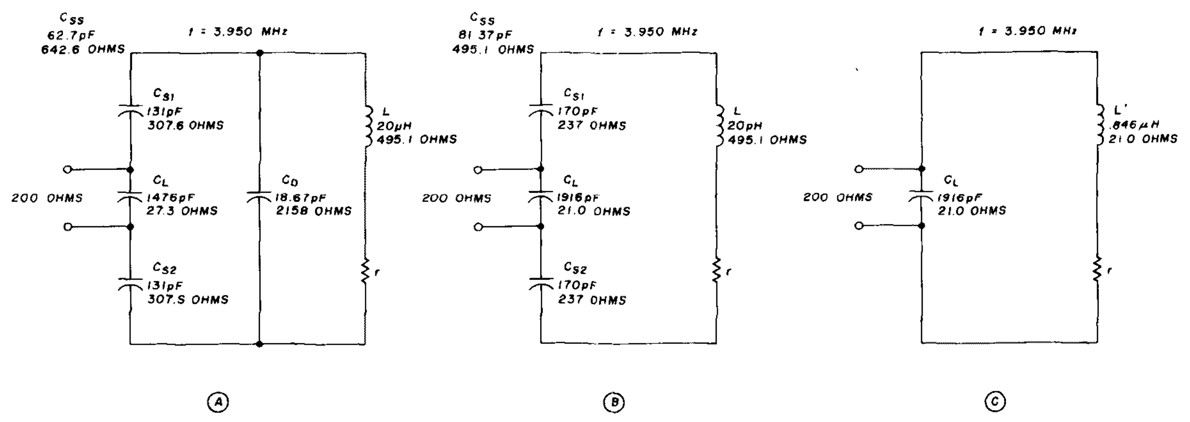

fig. 5. Transposition of the actual circuit of the SEED antenna (*A*); to the equivalent circuit (*B*); and to the simplified equivalent (*C*), as described in *appendix B*.

Maybe someone who reads this can publish information on how to economically obtain capacitors which can handle any of the following requirements:

7500 pF, 42 amperes, at	500 volts RMS, at 2 MHz
680 pF, 42 amperes, at	4,700 volts RMS, at 2 MHz
100 pF, 24 amperes, at	5,600 volts RMS, at 4 MHz

appendix A

With reference to **fig. 3A**, the parallel tuned circuit consists of an inductance, L, a distributed capacitance, C_D, and a series combination of tuning and loading capacitors, C_{SS}. The values across C_{SS} were measured as follows:

1.900 MHz	333 pF
3.950 MHz	62.7 pF

In each case, the distributed capacity must be added to that measured to resonate with L.

The formula for resonance,

$$f = \frac{1}{2\pi\sqrt{LC}}$$

can be rewritten:

$$\frac{\left(\frac{1}{2\pi f}\right)^2}{C} = L$$

Substituting appropriate numbers,

$$\frac{\left(\frac{1}{6.28 \times 1.9 \times 10^6}\right)^2}{333 \times 10^{-12} + C_D} = L$$

$$= \frac{\left(\frac{1}{6.28 \times 3.95 \times 10^6}\right)^2}{62.7 \times 10^{-12} + C_D}$$

which can be solved for C_D (18.67 pF) and then for L:

At 1.900 MHz: $333\ pF + 18.67\ pF$
$= 351.67\ pF$ or $238.2\ ohms$
and $238.2\ ohms = 19.953\ \mu H$ for L
at 3.950 MHz: $62.7\ pF + 18.67\ pF$
$= 81.37\ pF$ or $495.2\ ohms$
and $495.2\ ohms = 19.952\ \mu H$ for L

appendix B

With the antenna resonant at 3.950 MHz, the value of the loading capacitor was adjusted for 50 + j0 ohms at the end of about a half wavelength of new RG 213/U cable. Subject to any imperfections in the balun, this indicated an impedance of 200 + j0 ohms across the loading capacitor. The measured value of this capacitor was 1476 pF for a reactance of 27.3 ohms. The reactance across the series combination of C_{S1}, C_{S2}, and C_L, or C_{SS}, was 643 ohms.

The ratio of X_{CSS} to X_{CL} is 643 to 27.3 or 23.553 to 1. The impedance ratio is the reactance squared, or 554.75 to 1. Therefore, the impedance across C_{SS}, at resonance, is the impedance across C_L, 200 ohms, multiplied by 554.75 or 110,950 ohms. This is the impedance across the parallel tuned circuit and, at resonance, it is purely resistive. Since the relationship of the series resistance, r, to the parallel resistance, R, is:

$$r = \frac{X^2}{R}$$

where X is the reactance of either the inductor or total capacitance of $C_{SS} + C_D$, or 495.1 ohms. Then

$$r = \frac{245,124}{110,950} = 2.2093\ ohms$$

or the total series resistance of the antenna is 2.21 ohms.

Alternatively, the circuit may be considered by transposing it as shown in **fig. 5**. The circuit was modified and measured at 3.950 MHz, as shown in **fig. 5A**. In the absence of C_D, the total of C_D and C_{SS} would need to be combined to resonate at the same frequency. Retaining the same ratio between C_S and C_L, the values can be computed and would be as given in **fig. 5B**.

It can then be seen that the net reactance of C_{S1}, L, C_{S2}, in series, is inductive and equal to the capacitive reactance of C_L. Considering this combination, L', the circuit can be redrawn as in **fig. 5C**. This is a simple parallel tuned circuit, and it is known that its parallel impedance at resonance is 200 + j0 ohms and may be expressed as R. Then, since the equivalent series resistance, r, is given by $r = \frac{X^2}{R}$ and X is the reactance of C_L, which is 21 ohms. Then $r = \frac{441}{200} = 2.205\ ohms.$

appendix C

The loaded Q of the antenna at 3.950 MHz can be found by dividing the reactance of L by the series resistance. Then $Q = \frac{495}{2.21} = 224$. It can be shown that for a Q of 224, the bandwidth for an SWR of 2:1, would be 13.225 kHz at this frequency.

Tests of the SEED at 3.800 kHz, where the coax was electrically a half wavelength long, measured the bandwidth for a SWR of 2:1 of 13.950 MHz. This tends to corroborate other tests and computations.

ham radio

electromagnetic interference
and the digital era

Component selection, grounding and shielding solve common problem of digital design

Whatever happened to the good old days when unwanted radiation consisted of RF on the microphone cord or key line and relative output power was measured by the length of the arc drawn from the final caps to chassis? A high wattage soldering iron and a handful of large disk capacitors was always enough to solve any floating RF problems and the pulsing of the gas filled tubes served only as an "on-the-air" indicator.

As more electronic consumer equipment entered the home, Amateurs found it necessary to clean up their signals by adding shielding and lowpass filters to prevent interference with home entertainment products. New terms such as radio frequency interference (RFI), electromagnetic compatibility (EMC), and television interference (TVI) became more common and Amateurs were forced to become knowledgeable about quarter-wave traps, hi-fi speaker bypassing, and other methods of preventing their signal from being received by their neighbor's equipment.

With the advent of home computers, the interference problem took a new turn. Now manufacturers were producing and mass marketing equipment capable of causing interference in radio and television reception. It didn't take the FCC long to determine that some new regulations were in order. The result of the FCC rulemaking was Docket 20780, which placed certain limits on both conducted and radiated interference, and Docket 80284, which outlined the testing requirements necessary to show compliance with 20780.

what are the requirements?

The FCC made two major decisions which set the schedule and defined technical requirements for industry. The first cutoff date, January 1, 1981, applied to personal computers and electronic games. All such equipment manufactured after this date was required to meet the new specifications. After October 1, 1981, all other subsequently produced computing equipment had to meet the new specifications, and after October 1, 1983, no devices could be sold, regardless of the date of manufacture, unless they were tested and certified to operate within the allowable interference limits.

Docket 20780 defines two different classes of equipment and specifies radiated and conducted limits for each. *Class A* includes commercial equipment, while *Class B* limits apply to home equipment. **Table 1** outlines the allowable limits for both classes in the two interference categories.

There are, obviously, some questions which the manufacturer must settle before he will know if the product he makes falls under the new regulations. First of all, just what is classified as computing equipment and what is not? According to the definition in Docket 20780, any device that generates or uses signals or pulses in excess of 10 kHz is designated a computing device. Secondly, if the manufacturer produces a device which is used in both residences and businesses, which specification must be met? I don't know the legal ramifications of this question, but if a sizable portion of the market is for home use, I suspect the Class B specifications will have to be met. So, is your new Hi-Tech DX transceiver covered by this regulation? Well, if it contains or uses digital circuitry with clock frequencies in excess of 10 kHz; memories, a CRT, or a switching power supply; disks, a tape-drive, printer, communications interface, or microprocessor, it probably *is*.

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why the sudden interest?

Life in the analog lane was almost always predictable, with any interference problems easily isolated to a particular frequency and circuit. However, digital electronics has overwhelmed the RF environment with binary clocks that produce harmonics into the Gigahertz range, plastic equipment enclosures, wall plug power supplies, and unshielded ribbon cable that acts as an antenna to radiate both the signals it is meant to carry as well as any other signals near the circuit connection.

By the very nature of the binary format, digital pulses are rich in harmonics. As the signal switches from the minimum circuit value to the maximum, the potential interference at any frequency is dependent upon the waveform characteristics. **Fig. 1** shows the parameters for a typical pulse waveform. If we assign values to these parameters for purposes of illustration:

$A = 4$ *volts*

$t_r = 5$ *nanoseconds* $= 0.005$ *microseconds*

$t = 0.5$ *microseconds*

$T = 1$ *microsecond* ($PRF = 1$ *MHz*; $PRF = 1/T$)

PRF (Pulse Repetition Frequency) $= 1/T$

Then calculate:

$A(t+t_r) = 4(0.5+0.005) = 2.02$ *microsecond-volts*

$A/t_r = 4/0.005 = 800$ *volts per microsecond*

These values can then be plotted on a standard conducted interference graph for the particular waveform used. The resulting interference level for the above waveform parameters is shown in **fig. 2(A)**. Now let's increase the rise time by a factor of 100:

$t_r = 500$ *nanoseconds* $= 0.5$ *microseconds*

$A(t+t_r) = 4(0.5+0.5) = 4$ *microsecond-volts*

$A/t_r = 4/0.5 = 8$ *volts per microsecond*

table 1. Conducted and radiated interference limits.

conducted (MHz)	class A (μv)	class B (μV)
0.45 – 1.6	1000	250
1.6 – 30.0	3000	250

radiated (MHz)	class A (μV/m)*	class B (μV/m)**
30 – 88	30	100
88 – 216	50	150
216 – 1000	70	200

*measured at 30 meters
**measured at 3 meters

These values are plotted in **fig. 2(B)**. As a final test, let's change the repetition or clock rate to 500 kHz in the first example:

$T = 2$ *microseconds* ($PRF = 0.5$ *MHz*; $PRF = 1/T$)

$t = 1$ *microsecond*

$A(t+t_r) = 4(1+0.005) = 4.02$ *microsecond-volts*

These values are shown as curve *C* in **fig. 2**. From the interference plots, it is obvious that the lower clock frequency produces slightly greater amplitude, while a decrease in rise/fall time produces less interference at the higher frequency ranges. The points marked as f_1 denote the point where the envelope begins to drop

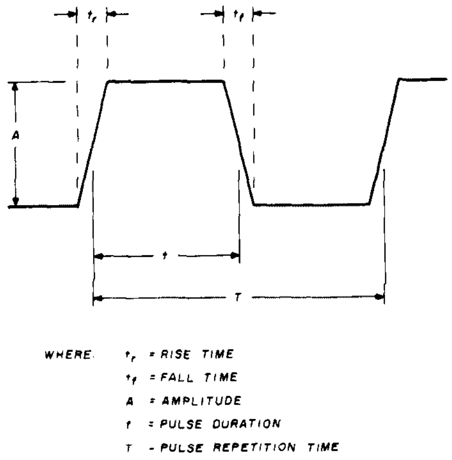


fig. 1. Pulse waveform parameters.

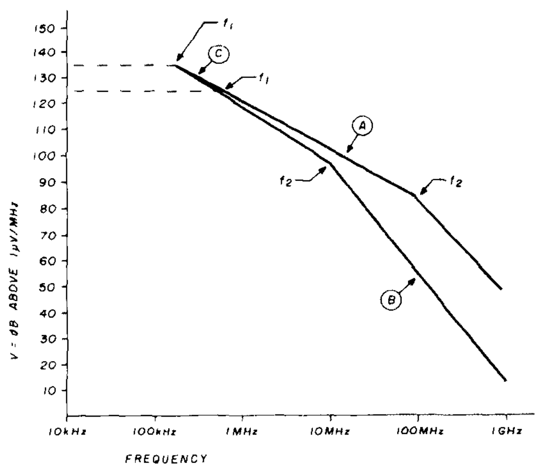


fig. 2. Conducted interference from symmetrical trapezoid.

off at the rate of 20 dB per decade. This point is determined by the pulse duration or period. Points shown as f_2 mark the frequency where the envelope drops off at a rate of 40 dB per decade. These points are determined by the rise and fall times. From this interference plot, it becomes clear that pulse shape, controlled by the rise and fall times, is the most important factor in reduction of the interference spectrum. This is further illustrated by **fig. 3**, which shows the relative interference levels of various waveshapes. It also becomes evident why operating a computer with poor electromagnetic interference protection precludes simultaneous use of everything from HF receivers and TV to the scanner.

conducted versus radiated interference

If we compare these two types of interference to a hot water heating system, the analogy may help us to understand the problem. The hot water radiator disperses heat to the room in proportion to the temperature of the water circulating through the system. If the water is not conducted to a particular radiator, no heat can be radiated from that radiator. Similarly, the interfering signals must be conducted from the source by means of circuit board traces, wiring, or components from which the interference can be radiated to a distance dependent upon signal strength.

Probably the only areas where conducted emissions are more important than radiated emissions are in the power line and grounding circuit. These areas become particularly important when switching type power supplies are used. With common power units, a multisection pi-type filter placed in a metal box, with *RF* continuity to the shielded equipment enclosure, will prevent any signals generated in the circuitry from reaching the power line. This procedure will also protect against interference radiation by the power cable. Conductive interference caused by ineffective grounding circuits is caused by an unintended resistive circuit ground connection to true ground. Such a connection will cause an offset signal level from true ground to be circulated to other points of the circuit which are also attached to the common bus.

Radiated interference commonly refers to any interfering signal propagated via an electromagnetic field. This definition includes radiation from wires or cables acting as transmitting antennas and coupling by mutual inductance or capacitance. The electromagnetic field propagates in normal fashion where the strength of the signal is inversely proportional to the distance from the source.

Interference contained on a signal cable between two pieces of equipment is not normally considered a potential radiated emission problem unless the conducted interference affects the operation of the unit to which it is connected.

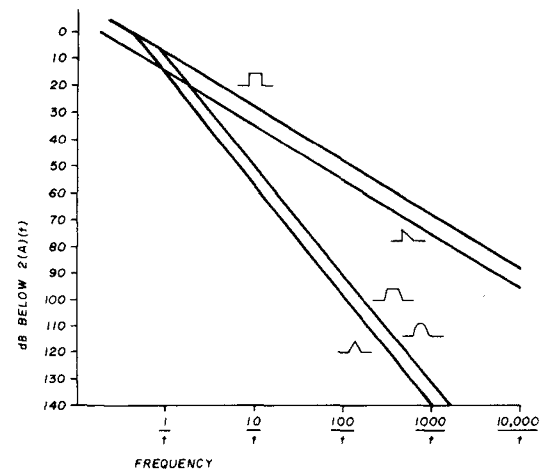


fig. 3. Interference levels for various waveforms.

design considerations

One of the worst jobs in the industry is trying to modify a piece of equipment manufactured without regard to EMC/RFI requirements to meet FCC regulations. In most cases, it is more economical to redesign, at least at the board or module level.

Good design practice dictates that EMI should be reduced as much as possible at every level. For circuit design, that means we must choose logic families with no greater bandwidth than necessary. For instance, CMOS has a lower bandwidth capability than other logic families. If a high speed clock is not required, use a lower speed clock instead of dividing down. Use waveforms with as long a rise and fall time and duration as possible within timing constraints.

When laying out the circuit board, use a minimum of one-eighth inch wide ground/common traces and place them at the edges of small boards, with additional traces down the center for larger boards. Ground traces should be connected at one end of the board only — the connector end. Ground returns should be as short as possible. If double-sided boards are used, interconnections between top and bottom surfaces should be frequent. Use of wiring to connect integrated circuits on different parts of the board should be avoided. When unavoidable, shielded cable, or at a minimum twisted pairs, should be used. Do not "float" any unused IC pins. Circuits employing clock frequencies on the order of 10 MHz and above are good candidates for double-sided boards. This procedure also acts to shield parallel boards in a vertical card slot configuration.

Liberal use of bypass capacitors will help minimize stray emissions. When strategically placed in the circuit, they will do the job even better. **Figure 4** shows the placement of various types and values of capacitors to suppress the *complete range* of frequencies.

Circuit board connectors present one of two locations where all signals are in close proximity to each other. If ribbon cable is used to parallel connectors, it should be mounted flat to the chassis and not run in close proximity and parallel to the component side of a circuit board. If various timing signals are to be carried on the same ribbon cable, alternate strands should be at ground potential. When using clock frequencies above 1 MHz, coaxial board connectors and cable should be considered.

The other location where the signal cables are in close proximity to each other is in the output connectors. If the output is a bus connection, it should be actively terminated rather than left to "float." This is one of the worst areas for emissions. If the outputs will not be continuously covered by an external connector, they must be covered by a shield (or metal shell) that prevents any leakage of RFI. External connectors must be metallic and use cables employing an overall shield that is continuous with the connector. For particularly bad cases, special chassis-mount connectors that incorporate filters as an integral part of each pin are manufactured. They operate much as DC feedthrough capacitors, as shown in **fig. 5**. Keep in mind that any filtering on signal lines will alter the waveform, much as a long length of cable would.

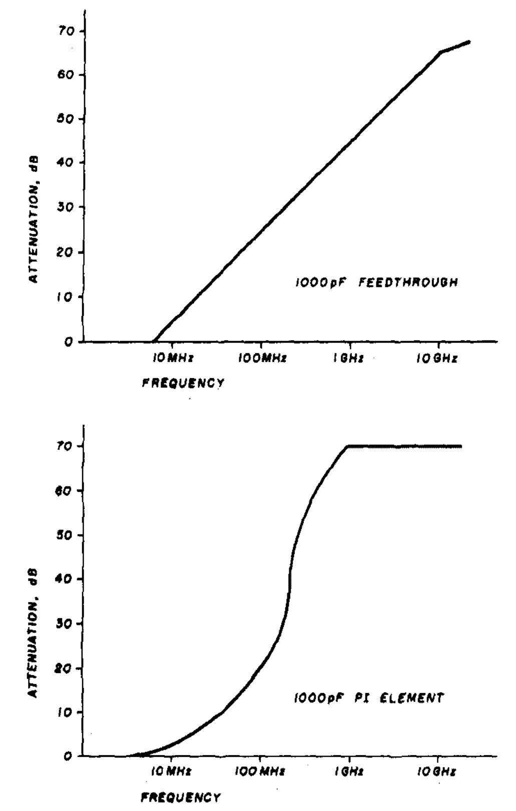


fig. 5. Typical attenuation for connector pin filters.

Therefore, if signal line filters are required, they must be selected by the effect they may have on the pulse shape, typically lengthening rise and fall times, but not necessarily by equal increments. If all else fails, fiber optics or optocouplers may be used.

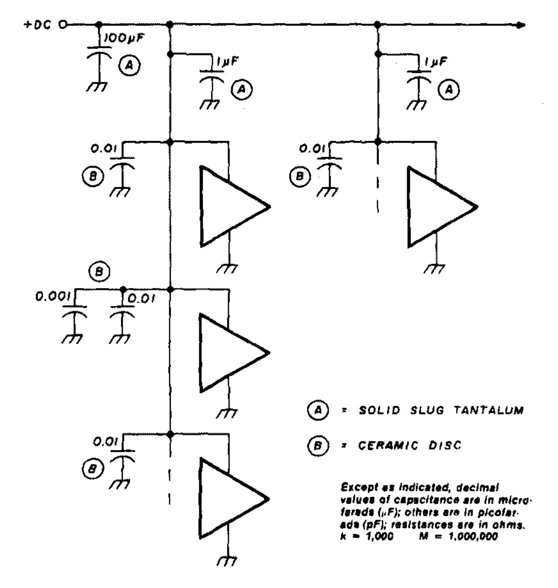


fig. 4. Bypassing for EMI suppression.

grounding

Although no guide to a perfect grounding system exists, practical experience has shown that following certain standard procedures is the first step in establishing an effective grounding system.

If you have control of that part of the ground system to which the equipment being designed will be connected (such as in a home or business environment), you should consider this as an integral part of the overall system. If a ground system is to act as a common voltage reference point, it is important that a single point, low potential, true ground location be established and that all branches feeding this point are low resistance/impedance paths. A typical system configuration is shown in **fig. 6**. Within each piece of equipment, all circuit boards/modules should have only one ground lead to the internal ground bus; if such a lead would be too long (over one-quarter wave-

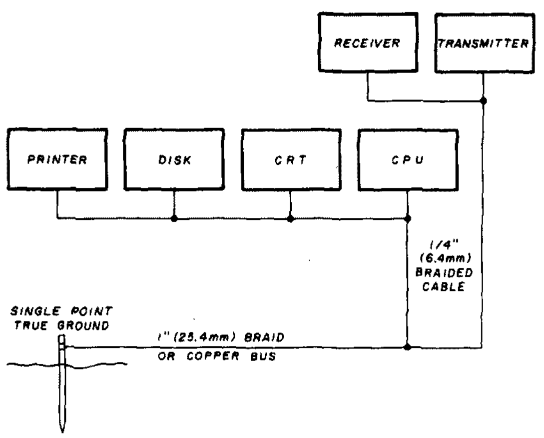


fig. 6. Representative single-point ground system.

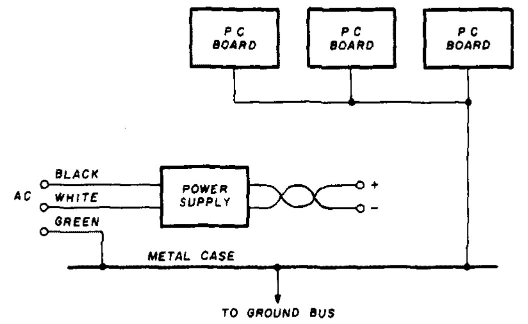


fig. 7. Internal equipment grounding.

length at the clock frequency), this lead may be connected directly to the chassis, as shown in **fig. 7**. Additional pieces of equipment comprising part of a system — i.e., CPU, disks, CRT, and printer — normally receive ground continuity via shielded cables and no additional ground wire connection between the chassis should be made.

shielding

Shielding refers to the use of an electromagnetic barrier to separate electric or magnetic fields. When implemented as an equipment enclosure, with appropriate measures to secure any openings against discontinuities, it functions to maintain all signals generated within the case as well as keeping potential external interfering signals from entering the enclosure. The enclosure material may be any one of various metals, a mixture of metal elements, or a plastic impregnated with metal bits or coated with metallic paint.*

table 2. Conductivity and permeability relative to copper.

metal	conductivity	permeability (150 kHz)	penetration loss dB/mil
copper	1.00	1	1.29
aluminum	0.61	1	1.01
brass	0.26	1	0.66
tin	0.15	1	0.50
steel (SAE1045)	0.10	1000	12.90

When the electromagnetic wave impinges upon the enclosure surface, it is not completely reflected. Although part of the energy is, in fact, reflected, the balance of the energy is transmitted through the material with the degree of attenuation (absorption) depending upon the type of enclosure material, thickness of the material, and frequency of the electromagnetic energy. (If we were to consider the shield as a plane of glass through which we are attempting to shine a light, perhaps this analogy would clarify the concept. The glass will reflect some of the light, depending upon the angle of the beam, but some will also be transmitted to the other side, with the amount transmitted dependent upon the thickness and color of the glass. In addition, there will be secondary and higher orders of reflections from each side of the glass plane of lower amplitude, as shown in **fig. 8**. For practical purposes, we may neglect the secondary and higher functions since they are mainly applicable to magnetic fields, and state that the shielding effectiveness is the sum of the reflection and absorption losses.)

In the selection of a suitable material for shielded enclosures, we must know the attenuation required, the frequency range of the potential interference, and the limits of thickness of the material (10 gauge steel

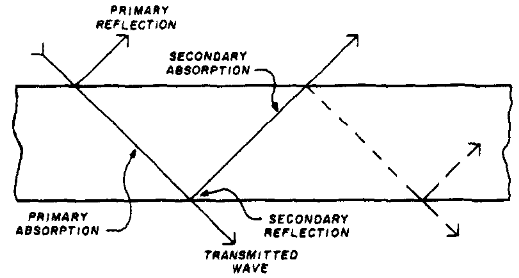


fig. 8. Reflection and absorption of electromagnetic wave.

*See Vaughn Martin's "EMI/RFI Shielding: New Techniques," *ham radio*, January and February, 1984.

table 3. Metal thickness required for 60 dB absorption loss at 1 MHz.

metal	thickness
copper	15 mils
aluminum	20 mils
brass	25 mils
tin	35 mils
steel	1 mil

table 4. Plane wave reflection loss at 1 MHz.

metal	reflection loss
copper	108 dB
aluminum	104 dB
brass	102 dB
tin	100 dB
steel	77 dB

does not lend itself to easy preparation). Generally speaking, ferrous metals are more effective shields at very low frequencies than nonferrous materials; sheet steel has medium effectiveness at these frequencies; nonferrous metals and steel are suitable at higher frequencies. The thickness of the shielding material is a function of permeability and can be found by:

$$t = \frac{A}{3.338 \times 10^{-3}\sqrt{fG\mu}} \ mils$$

where A = required attentuation
f = frequency in Hertz of lowest interfering signal
G = conductivity of shielding material chosen compared to copper
μ = the relative permeability of the shielding material

Conductivity and permeability values for various metals are given in **table 2**. Choosing the wrong material may require a shield thickness that is impractical. The thickness required for various metals to provide 60 dB absorption attenuation at 1 MHz is shown in **table 3**. For reflection loss, **table 4** lists the plane wave attenuation for the same metals at a frequency of 1 MHz. A tradeoff of skin thickness versus attenuation required can be made in cases where one area of circuitry is producing the strongest interference. By installing a shield around the offending components on the circuit board, the requirements for enclosure shielding are lessened.

Holes in the case for ventilation, connectors, switches, etc., will materially decrease the shielding effectiveness. The larger the diameter of the hole and the higher the frequency, the greater the leakage. Ventilation openings may be covered with wire mesh, but it must be directly bonded to the case for maximum effectiveness. Preformed metal cases with hole patterns will decrease the attenuation in proportion to the hole diameter, depth, and size of the area containing the openings. Although the cutoff frequency of the openings can be determined from the diameter

$$f_c = \frac{6.92}{d}$$

$$f_c = 1 \times 10^6 = \frac{6.92}{d}$$

solving for d, we obtain $d = 0.000007 \ inches$

the depth or thickness of the material must be three times the diameter of the hole to produce 100 dB attenuation.

Where parts of the shielding enclosure must be bolted together, the opportunity for leakage is great. For example, with perfectly flush, bare metal contact between two pieces of one-half inch, 0.09 inch thick aluminum, shielding effectiveness decreases by 25 dB when mounting hole spacing is increased from one inch to five inches. Using woven knitted wire mesh gasket material in these areas, as well as around connectors, will maintain the shielding integrity of the enclosure. As a general rule of thumb, where RF gasketing is not used, or the enclosure contains discontinuities such as corner bend strain relief openings, multiple bolt-together sections, openings for switches, fuses, etc., the maximum expected attenuation is about 30 dB.

The same type of rationale applies to the use of coax cable. Leakage through the braided shield depends on the material used, the number of strands used, and the spacing or turns per inch. Cable is available with various values of shielding effectiveness to 95 percent for double shielding. The shielding effectiveness of coax cable is *only as good as the connectors used at each end.* A double-shielded cable yielding 95 percent effectiveness with 60 dB connectors will not provide more than 60 dB attenuation.

conclusion

Although we can estimate the required shielding for a given attenuation specification, it is impossible to forecast the interference level of the circuitry in the design stage. Therefore, it is imperative that procedures to minimize RFI be implemented at the circuit board level to lessen the system shielding requirements.

Now that Amateurs are more involved in designing their own digital circuitry for use with their other station equipment, we must put appropriate EMC procedures into practice or risk operational degradation to the companion units or even loss of use of the complete station because of RFI.

bibliography

Electromagnetic Compatibility Manual, NAVAIR 5335: NTIS AD-754 411 National Technical Information Service, United States Department of Commerce, 5285 Post Royal Road, Springfield, Virginia 22151.

ham radio

solving the problems of RFI

A brief recollection of quirks and cures

Whenever there's an RFI problem with a piece of equipment, the instruction manual inevitably suggests connecting the equipment to a good ground. This is appropriate advice if the equipment is all in the basement, or at worst, on the first floor. But if your gear is spread around the house, obtaining a good ground at Amateur Radio frequencies may not be easy.

While some grounds *are* more effective than others, sometimes it's better not to use a ground at all, but to use other means — such as filters — instead. Those of you who've worked in broadcasting, and are familiar with the sensitive, low-level audio circuits found in studios close to the transmitting antenna, know that it's a common practice to ground only the shield at the input end of the audio circuit and leave the output end ungrounded. This prevents a ground loop that can aggravate a hum problem rather than alleviate it.

It's a mistake to tie everything to one ground lead. The more equipment connected to a ground lead, the greater the resulting current flow, creating more problems with interference. Better to keep all grounds, especially AC and RF grounds, separate. The feedthrough of RF power to the AC line will be much greater from a common junction than the AC line would pick up from direct radiation.

A classic case of RF power being fed through a common junction was a commercially built one-kilowatt RF amplifier designed for Amateur use. Probing around with an RF meter fitted with a figure-eight pickup loop showed plenty of RF current flowing in both the line cord and the house wiring. This was rather puzzling because the amplifier has a built-in AC line filter of good construction, consisting of sufficient turns of heavy wire on a half-inch ferrite rod, five inches long.

The amplifier had a built-in power supply with dual windings on the power transformer primary so that it could be operated from either a 115 volt or 230 volt power line. Finally the words, "the neutral (green) wire of the line cord should be connected to the chassis at all times" caught my attention.

That phrase gave me the answer — an answer that for some reason, escaped the attention of the design engineer. The direct connection of the neutral wire to the chassis and to the AC ground nullified any filtering action of the AC line filter. The cure was obvious.

If the amplifier had been operating on 230 volts, I would have had to wind a third coil over the line filter, or if that were too difficult, to wind a separate and suitable RF choke and connect it between the chassis and the neutral lead. Since I was operating the amplifier from the 115 volt line, all I had to do was to disconnect the neutral (green) lead from the chassis and tape it. This reduced the RF current flow in the cord and the house wiring to a barely noticeable flicker of the needle on the RF meter.

When bypassing flickering lights, "hot" outlet receptacles, transformers, and/or motors, instead of following the customary method of connecting a capacitor from each side to ground, I connect a capacitor across the line and then connect another capacitor from the hot side to ground.

This practice began when I installed a large fluorescent light above my workbench. It gave very good light but it also put out a lot of RF "hash," making it impossible to do any signal tracing or alignment work with the light on. The fixture had a ballast-type coil or transformer in the case and evidently connecting the bypass capacitors in the usual manner made the capacitors act like a voltage divider. Connecting the capacitors the second way cleared up the "hash."

This brings to mind a home-made power RF amplifier that was very unstable. Neutralizing it on 10 meters made it unstable on 20 or 40. The suggested cure for such a condition is to try to achieve a low impedance path from the filament to ground by adding capacitors in parallel, but doing so only made things worse. The cure was to *reverse* my efforts.

By John Labaj, W2YW, 12 Park Place, Elsmere, New York 12054

Instead of loading more and more capacitors on the filament lead I removed all capacitors from the filament lugs and wired a small RF choke consisting of eight turns of No. 14 wire wound on a 2-watt 220-ohm carbon resistor in each filament lead directly at the socket. The other end of the choke that was connected to the filament transformer was bypassed with just one 0.001 μF capacitor.

The deliberate introduction of some negative feedback made the amplifier stable. The neutralization was effective on all bands. There was also a marked reduction in the amplitude of the harmonics coming out of the amplifier. The only change from normal was that the amplifier required a little more drive for full output on 10 meters.

When the interference cure you install does not provide the expected result, don't assume your idea wasn't good; perhaps the item or items you used simply failed to do the intended job. This was vividly demonstrated to me early in my struggle with television interference when I decided to feed all the television sets in the house through coaxial cable from a distribution amplifier. The results did not meet my expectations.

Noticing that some parts varied in performance, I made a deal with a helpful local distributor. He allowed me to charge out every make of every item that I'd need to do the job with the understanding that I could return for full credit anything I did not use. What a revelation!

Sometimes an item made by a well-known company gave very poor or even negative results, while the same item made by a lesser known company did a much better job. This applied to such items as distribution amplifiers, high-pass filters, matching transformers and AC line filters. So if your first choice doesn't work, don't give up; try using the same part, but from a different manufacturer.

In summary, remember that just because the wire is connected to earth ground doesn't mean that it's an effective ground at all frequencies. Run *another* ground lead even if it seems to be a needless duplication.

Keep AC ground and RF grounds separate. Also, if the cure you try doesn't work, don't remove it right away. Many times a combination of two or more remedies will be effective where the single cure is not.

It's been my observation that home-made AC line filters work better than the commercial variety because they can be made to fit the load and space. Low-pass filters never worked for me; they seemed to waste power, and if you cranked up the power, they went up in smoke. Use coax stubs cut for the affected channel.

If matching transformers don't have insulated windings, modify them by rewinding.

ham radio

book and product

REVIEWS

Computer Programs for Amateur Radio

by Wayne Overbeck, N6NB, and James Steffan, KC6A

Amateur Radio is an "information intensive" avocation. Detailed and accurate records are imperative if you're working toward WAS, DXCC, or VUCC awards. Contesting, too, requires that accurate original and "dupe" logs be kept. And if you're a DXer, you also want to know such things as beam headings, distances to DX stations, sunrise and sunset times for grayline propagation, and much more. All this data can be stored in a home computer and retrieved with ease. But many hams use their home computers for nothing more than video games, RTTY, and Morse code applications because they don't have the programming skills to do more. No wonder — until now, there haven't been many programs available for Amateur Radio applications. Those that were available were usually written for just one machine and couldn't be used with others without extensive revision.

The authors of *Computer Programs for Amateur Radio* — who have over 50 years of Amateur Radio experience in all phases of operation, from contesting to electronic design — were aware of this phenomenon and set about the task of writing progams for just about every aspect of radio operation. The programs will work with most popular home computers.

Chapter One gives the reader an overview of the programs contained in the book. Programs were selected according to two basic criteria: first, that they be useful in the hamshack and fully tested, and second, that they be compatible with the most popular brands of home computers. They can be used with the Apple II, IBM PC, TRS-80, Commodore C-64, and any other CP/M and Microsoft BASIC unit. Owners of the VIC-20 and Timex/Sinclair or Sinclair units will find that a number of the programs will also run on their machines.

You'll need at least 48K of memory and a disk drive. A printer is optional but highly recommended. Chapter One also provides a listing and explanation for each program in the book. Programs include *Logbook, Radio Awards Data Base, Grid Locator, Worldwide "Catalog File," Sunrise Chart, Sunrise Calculator, Grayline, Beamheading Chart, DX Display, DX Checker, Dupe Checker, Dupe Print, Contest Logger, Generalized Logger, Field Day Logger, Sweepstakes Logger, Log Print, Antenna Scaler, Antenna Evaluator, Phased Vertical Pattern Plotter, EME System Analyzer, Moontracker,* and *Skylocator.*

Now that you know what you're going to get, are you ready to use it all? Anticipating that readers would need additional help, the authors devoted Chapters Two and Three to a discussion of microcomputers in general, including their evolution and associated hardware. Chapter Four provides an overview of some of the problems often encountered; it's not meant to discourage, but rather to prepare the user for the potential problems that can and will occur. Chapter Five is a rather complete discussion of BASIC (*B*eginners *A*ll-purpose *S*ymbolic *I*nstruction *C*ode) included for the purpose of providing a solid grounding in the most frequently used computer language. While BASIC is slower than assembly language, it is quite flexible and easy to write and revise. To illustrate how BASIC is used, the authors use a "Mini Logger" program as an example. A flow chart and full step-by-step description of the program assist the reader in learning the program.

The rest of the book, Chapters Six through Ten, are devoted specific programs. Each program is fully discussed, documented, and presented with hints to help the reader obtain maximum use of each.

The authors have gone to great lengths to provide all the information necessary to help readers get the most out of their home computers. Beginners and experts alike will find this book full of helpful information.

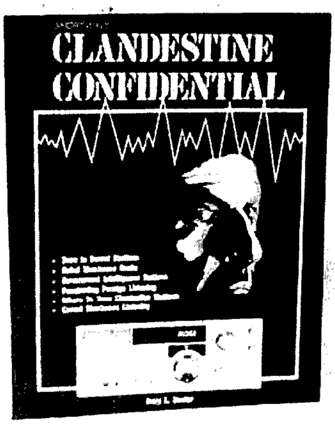
Computer Programs for Amateur Radio is available from Ham Radio's Bookstore, Greenville, New Hampshire 03048, for $16.95 plus $2.50 shipping and handling.

Circle #301 Reader Service Card.

NIACH

Clandestine Confidential

A new book published by Universal Electronics, Inc., covers the world of clandestine broadcasting yesterday and today. Subject areas covered include clandestine DX'ing, the history of clandestines, a listing of 30 active countries with their frequencies, discussion of where new clandestines may surface in the future, and clandestine QSLing. A list of clandestine station addresses is also provided.



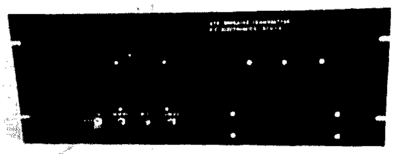
The retail price of *Clandestine Confidential* is $8.95 plus $1.75 shipping and handling in the U.S. and Canada.

For further information, contact Universal Electronics, Inc., 4555 Groves Road, Suite 3, Columbus, Ohio 43232.

Circle #321 on Reader Service Card.

ATV repeater transmitter

A new 40-watt PEP ATV repeater transmitter is available from P.C. Electronics. The 7 × 19-inch rack panel RTX-4 transmitter comes crystalled for the normal ATV repeater output frequency of 421.25 MHz, but can be ordered for any other frequency in the 70 cm Amateur band between 420 and 440 MHz for transmitting weather watch or other emergency service bulletins, NASA space shuttle video, or even beacon and base station use.



The transmitter accepts the normal 1-volt composite video, either color or black and white, and mike or line audio. Besides the video output from a color TV/monitor normally used in the repeater application, any device with a composite video output, such as a camera, VCR, computer, TVRO, etc., can simply be plugged into the front panel jacks and transmitted.

The RTX-4 contains the VOR (video operated relay) module, which keys the transmitter on only when a video signal containing the normal horizontal sync frequency around 15.75 kHz is sensed at its video input. This prevents false keying from military radars, commercial radio positioning, and other Amateur modes that may be sharing the repeater input frequency range.

Power requirements are a regulated 13.8 VDC at 8 amperes and 120 VAC at 0.1 ampere for operation. Notes on how to successfully assemble a complete basic ATV repeater system for under $2000 and the technical considerations unique to ATV are included.

For more information on this ATV transmitter and other ATV products, contact P.C. Electronics, 2522 S. Paxson Lane, Arcadia, California 91006.

Circle #302 on Reader Service Card.

ham gear protection

The new 40-page catalog from Electronic Specialists includes a line of protection and interference control products designed to prevent costly damage from lightning or power line spikes and disruptions or interference from power line carried EMI and RFI. Protective devices include AC line voltage regulators and conditioners, modem and phone line surge suppressors as well as equipment isolators and filter/suppressors.

Typical protection and interference problems are described in the text, together with suggested solutions for various Amateur and other communications installations, as well as numerous applications for hi-tech equipment protection and interference control.

For more information, contact Electronic Specialists, Inc., 171 South Main Street, P.O. Box 389, Natick, Massachusetts 01760.

Circle #303 on Reader Service Card.

DMM/DCM

MCM Electronics has introduced the Temma Combination DMM/DCM meter with hFe transistor gain tester. The unit measures voltage, current, resistance, capacitance and hFe on the clear 1/2-inch, 3-1/2 digit LCD display.

A capacitance measuring socket gives direct measurements of capacitors, along with a transistor hFe. The color-coded panel allows users easy identification of function and range settings.

Safety features include input overload protection, single fusing (with a spare fuse inside), and stress relief test leads. The Temma Combination DMM/DCM meter comes in a convenient carrying case, with alligator clip hFe leads and has

NEW products

a one-year warranty. Battery operated, the LCD display indicates low battery condition. The price is $74.95.

For more information, contact MCM Electronics, Centerville, Ohio 45459.

Circle #304 on Reader Service Card.

ShackMaster™

A new product from Advanced Computer Controls, Inc., allows remote control of your shack and effective communication with family members over your home equipment.

ShackMaster's™ crossband linking capability allows you to access your high performance home station from VHF/UHF, either by simplex or through repeaters. Telephone access permits remote control of your home station from any Touch-Tone telephone, and BSR X-10 shack control offers Touch-Tone remote control of 120-volt devices with Touch-Tone commands, over the air or over the phone.

The ShackPatch™ feature, a remotely controlled intercom, permits remote control of your home equipment, allowing third parties to participate. Based on the same principles as an autopatch, it leaves you in complete control of your station at all times. An electronic mailbox permits you and your family to leave messages for each other, to be retrieved when convenient. Finally, a simplex autopatch is available when it's necessary to make a phone call.

Based on ACC's proven repeater control technology, ShackMaster™ includes electronic synthesized speech with a custom vocabulary, making it easy for anyone to use. It interfaces to up to three transceivers, the phone line, a local speaker, and microphone.

For more information, contact Advanced Computer Controls, Inc., 10816 Northridge Square, Cupertino, California 95014.

Circle #305 on Reader Service Card.

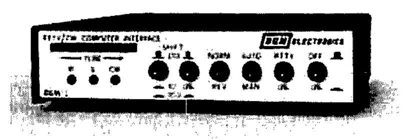
RTTY/CW computer interface

DGM Electronics, Inc., has introduced their new DGM-1 RTTY/CW computer interface, which simply connects between any transceiver and computer and works with almost any RTTY/CW software on the market because of its versatile I/O circuitry.

The RTTY demodulator provides strong performance even on the weakest, noisest signals that can be found. This is because of the sensitive mark and space active filter demodulator, rather than a phase-locked loop, as found in other low cost interfaces. This unit copies both the mark and space tones, not just the space tone. The demodulator section is preceded by a bandpass filter to provide excellent adjacent signal rejection. A three-pole post detection filter provides optimum signal-to-noise reception of the RTTY signal. The 170, 425, and 850 Hz shift selector provides fast and accurate shift selection; shift can also be reversed with the use of a front panel pushbutton. An LED bargraph and mark/space LED indicators provide positive tuning indication. Scope outputs are also provided for the ellipse tuning. A function generator chip is used to provide a stable, sinewave AFSK output to your transmitter. This interface will also key your FSK input. Automatic or manual PTT control can also be selected by a front panel pushbutton.

The CW demodulator, centered around 800 Hz, includes bandpass filtering to reject nearby signals. Both positive and negative CW keyed outputs are provided on the rear panel.

The rear panel contains a standard 5-pin I/O connector for TTL level interfacing. These signals can be inverted so that just about any software can be used with the interface. An RS 232 connection is also included for use with computers requiring these voltage levels.



The DGM-1 RTTY/CW computer interface is housed in a compact 1-1/2 × 7 × 7 inch aluminum enclosure to provide excellent RF immunity. The unit is powered by a 120 VAC wall transformer, included with the interface. The price of the DGM-1 is $149.00.

For more information, contact DGM Electronics, Inc., 787 Briar Lane, Beloit, Wisconsin 53511.

Circle #306 on Reader Service Card.

coax checker

North American SOAR Corp.'s Model 1500 coaxial cable length checker was designed to meet the needs of cable manufacturers, users, and installers.

The length of coax and its termination — whether on a reel or strung out — is essential information for anyone who works with it. Using the Model 1500 coax checker, damaged cable in plenum, conduit or in free air can be checked for shorts or opens. The device provides numerical indication to the short or open in feet or meters and indicates the state — either short

or open — on a 4-digit LCD readout. This unit can measure all types of coaxial cable in lengths ranging from 10 feet to 6500 feet. The pulse reflection technique used as its measuring method allows for fast cable length indication.

Totally portable, the unit measures only 7-3/8 × 2-1/4 × 7 inches and weighs only 3 pounds with batteries. Priced at $499 in small quantities, it is supplied with 6 Ni-Cad rechargeable batteries, a battery recharger/AC adaptor and a car/truck cigarette lighter adaptor.

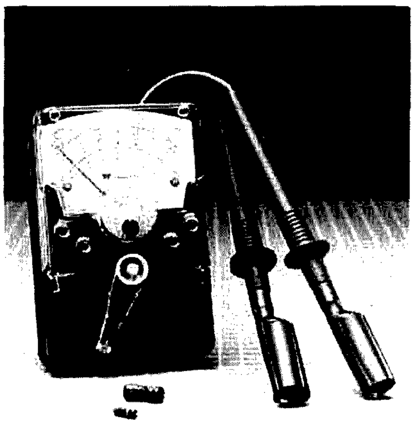
For more information, contact North American Soar Corp., 1126 Cornell Avenue, Cherry Hill, New Jersey 08002.

Circle #307 on Reader Service Card.

Triplett hand-size tester

The new Model 310-T5 hand-held V-O-M just introduced by Triplett Corporation offers an extended AC/DC voltage range up to 1200 volts for extra versatility in making laboratory or in-field measurements on industrial, commercial, or consumer electronic/electrical equipment.

The tester has a sealed range switch for improved resistance to contaminants and new, safety-designed test leads to provide optimum user safety. The drop-resistant case is high impact thermoplastic to endure rugged use.



Only 2-3/4 × 1-5/16 × 4-1/4 inches, the Model 310-T5 has an easy-reading, 2-1/8-inch scale meter which is self-shielding and is protected against overload. Full scale accuracy is ±3 percent DC and ±4 percent AC.

Ranges include: 0-1200 VDC and VAC in 5 ranges; 0-200 megohms in 4 ranges; 0-600 DC milliamperes in 4 ranges with a 0-600 microampere (250 mV) range. A convenient single range selector switch is provided.

Tester sensitivity is 20,000 ohms/volt DC and 5,000 ohms/volt AC.

Priced at only $70, the one-year warranteed Model 310-T5 tester is furnished complete with 42 inch safety test leads, screw-on alligator test clips, batteries and a comprehensive instruction manual. Optional accessories include the Model-10 clamp-on AC ammeter, Model 101 line

NEW products

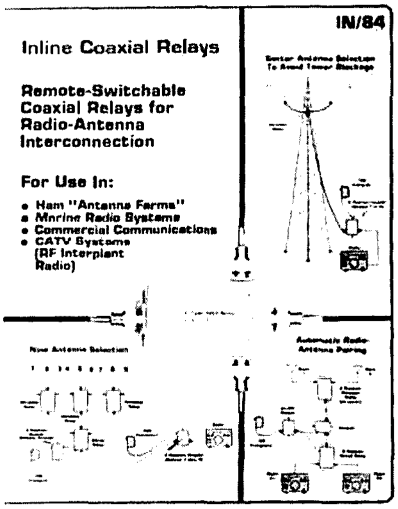
separator, carrying cases and replacement test leads.

For additional information, contact Triplett Corporation, One Triplett Drive, Bluffton, Ohio 45817.

Circle #308 Reader Service Card.

In-line coax relays

In-line coaxial relays can remotely select from up to nine antennas using a single coaxial cable to the radio.



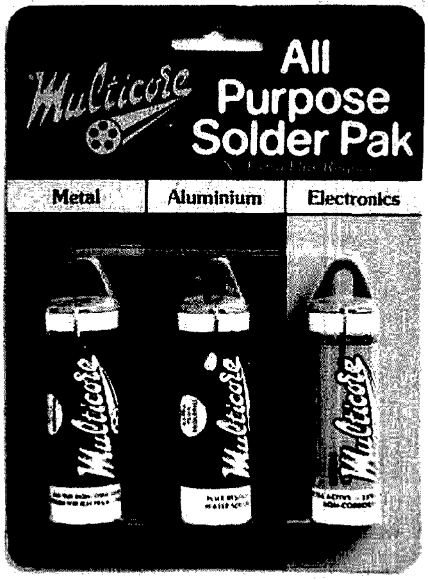
The two and three-output relays can be arranged in systems to select from among multiple antennas, pair different antenna-radio combinations, or connect a series of radios to a single broadband antenna. The various relays cover some part of the 0-900 MHz range, and feature high power handling, long life, and weatherization. These systems as relays, together with ancillary couplers, switches, and power supplies are described in catalog IN/84. They range in price from $40 to $66, depending on configuration.

For more information, contact Microwave Filter Company, Inc., 6743 Kinne Street, East Syracuse, New York 13057.

Circle #311 on Reader Service Card.

solder kit

A new selection of solders for virtually any type of soldering job — including metal, aluminum, and electrical wiring — is available from Multicore Solders. The "All Purpose Solder Pak" contains an assortment of multiple-core construction wire solder for self-fluxing action. The



correct flux formulation and alloy composition provide the ideal combination for different soldering jobs: metal, aluminum, or electrical/electronic. No pre-cleaning of surface is necessary. Each plastic dispenser has a convenient hook-eye cap for bench-top storage.

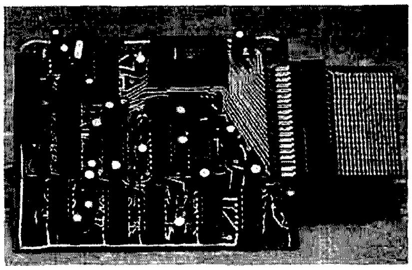
Included in the selection are Arax Solder for general metal repair, Alu Sol Solder for aluminum, and Ersin Solder for electrical/electronic connections.

For complete information, contact Multicore Solders, Cantiague Rock Road, Westbury, New York 11590.

Circle #310 on Reader Service Card.

Touch-Tone remote control board

TTC300, is a new DTMF (Touch-Tone) Controller Board that provides remote DTMF control of virtually any ON/OFF function via a radio or any type of link with audio output, such as wireline or phoneline. Typical applications include remote control of functions at a repeater site or any location with a radio link.



The controller uses a new high quality crystal controlled decoder IC, with high immunity to falsing, decodes all 16 digits, and features 3 ON/OFF functions per main card. Easily expandable to any number of functions with expan-

sion cards, the board can be field programmed using plug-in coded cards.

Its transistor switch outputs can directly trigger solid-state circuitry or relays for any type of control function.

For more details, contact Spectrum Communications Corp., 1055 W. Germantown Parkway, Norristown, Pennsylvania 19401-9616.

Circle #312 on Reader Service Card.

code teacher program

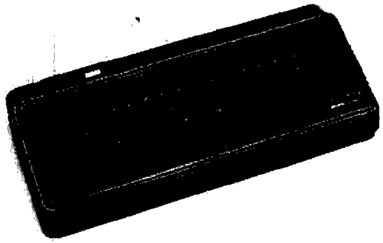
Cynwyn now offers MC-10 owners Morse Code Teacher, a program requiring 4K RAM. Available on cassette for $15 plus $2 shipping and handling, Morse Code Teacher is designed for the beginner. It features three different practice routines that promote familiarity with the code and can increase copying speed up to 5 WPM. In the introductory routine, whenever any letter or number on the computer keyboard is pressed, the program responds with the equivalent character in Morse Code. The second routine generates and sends characters one at a time from pre-determined letter/number groups and displays them on the screen for checking. In the final routine, random letters and numbers are sent at either 3 or 5 WPM for copying sessions of one minute and displayed on the screen at the end of each session.

For more information, contact Cynwyn, 4791 Broadway, Suite 2F, New York, New York 10034.

Circle #313 on Reader Service Card.

CW keyboard

The HD-8999 UltraPro CW Keyboard is a third-generation code computer designed to minimize keying errors and increase the ease and accuracy of sending high-speed CW. A 64-character "type ahead" buffer permits typing faster than the keyboard is sending. Ten variable length buffers



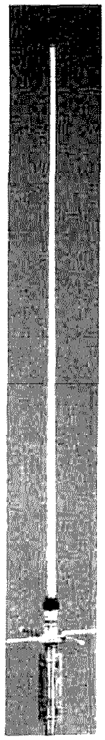
eliminate waste when storing text, and messages stored in the buffers can be compiled, corrected, or transmitted with no more than one to three keystrokes. A large, four-digit LED display indicates many functions including speed, spacing, weighting, serial number, remaining message character space, input error, tune mode, sidetone on/off, keyclick and individual buffer protection. An 8-segment bar graph indicates buffer protection. An 8-segment bar graph indicates fullness of the type-ahead buffer. Parameters are easily set from the keyboard,

NEW products

and battery back-up of the CMOS memory retains buffer contents and last-used parameters in the event of power failure or the keyboard's being turned off. Three different, four-level code practice modes are built in, as are turn-on circuit diagnostics, a sidetone oscillator, and speaker.

For more details, contact Heath Company, Benton Harbor, Michigan 49022.

Circle #309 on Reader Service Card.



UHF fixed station antenna

The new G6-440 UHF antenna for fixed station or repeater use was recently announced by Hustler.

Based on the popular G7-144 VHF antenna, the new UHF antenna delivers 6 dBd gain through the use of stacked 5/8-wave brass radiator sections, series phased, and sealed in an ultralight, tapered fiberglass radome.

Mechanical integrity is assured with the use of aluminum and stainless steel components. Coaxial cable termination is accomplished through the use of a hub-mounted, moisture resistant, "N" type connector.

The antenna is factory tuned at 440 MHz, with a typical VSWR of 1.15:1, and exhibits an 18 MHz bandwidth under 2:1. Its overall height is 88 inches, with a wind survival rating of 120 MPH.

For further information, contact Hustler, Inc., 3275 North B Avenue, Kissimmee, Florida 32758.

Circle #315 on Reader Service Card.

Touch-Tone decoder and encoder/decoder

Midian Electronics, Inc. has introduced the TTD-3 and TTC-3. The TTD-3 is priced at $59.95, and the TTC-3 is priced at $85. The TTD-3 is a 1 to 4-digit diode snip programmable anti-falsing DTMF decoder. It can decode, A,B,D,D,*,#, 1 0. It has a 2400-Hz ring tone, momentary horn output, latchng call light, and positive or negative squelch output. The unit measures 1.17 × 1.15 × 0.3 inches.

The TTC-3 is a combination encoder/decoder with all the features of the TTD-3 plus push-to-talk, sidetone audio to the speaker and adjustable audio output, and all the 16 standard Bell System touch tones.

For additional information, contact Midian Electronics, Inc., 2302 East 22nd Street, Tucson Arizona 85713.

Circle #314 on Reader Service Card.

code and theory tapes — in stereo

Gordon West's Radio School offers over 30 individual 1-1/2 hour long code speed-building courses on stereo cassettes. There are also over 20 individual tapes covering theory for examination preparation, and 10 tapes dealing with Amateur Radio equipment installation techniques.

West's stereo technique allows students to play the tapes in a variety of ways to satisfy their individual learning requirements. Any tape player with a balance control can be used to fade out the voice channel as needed. Played on a monaural tape recorder, the student hears both channels.

Radio School also offers complete 4-cassette theory courses covering the new FCC questions from Novice to Extra class. These theory courses feature the "live sounds" of Amateur Radio operating to assist the student in recognizing some of the topics discussed on the tape.

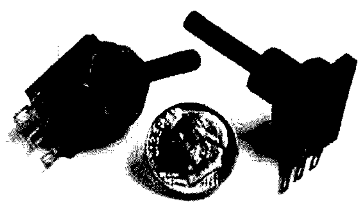
All Gordon West Radio School tapes are available directly from Radio School, 2414 College Drive, Costa Mesa, California 92626.

Circle #316 on Reader Service Card.

all-plastic potentiometers

Mouser Electronics has announced the release of a new hi-rel potentiometer said to have been designed for safety.

The 31N series pots are suitable for applications requiring both high voltage and high insulation resistance. Both body and shaft are made of flame-retardant nylon with a conductive plastic element. These pots can handle a working voltage up to 315 VAC (630 VAC peak) with a minimum insulation resistance of 1000 megohms.



They are linear taper and are power rated at 1/4 watts at 20 degrees C. Resistance tolerance is 20 percent and terminal resistance is a maximum of 5 ohms. Insulation voltage is 450 VDC. The pots measure 0.79 × 0.6 × 0.68 inch shaft.

Production quantities are available from stock in values from 1 K to 1 M and come complete

NEW products

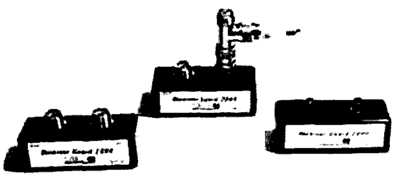
with hardware. Prices are as low as 99 cents in quantities of 500.

For further information, contact Mouser Electronics, 11433 Woodside Avenue, Santee, California 92071.

Circle #317 on Reader Service Card.

receiver guard

Design Electronics Ohio (DEO) has announced the Receiver Guard 2000, a solid-state, RF-triggered protection device that prevents high power RF from damaging modern solid-state front ends.



The unit may be installed in any receiver antenna line, or with slight modifications, in the transverter jack of several popular transceivers. Once installed, the unit is totally passive until RF voltage on the antenna exceeds 1 volt (1000 millivolts). At this threshold voltage, the unit begins to activate, shunting excess voltage to ground while automatically increasing the resistance in the line to the receiver. This automatic increase in receiver line resistance continues until a fusible link inside the Receiver Guard 2000 opens.

Many Amateur Radio applications are possible. This unit is suited for use in the multi-transmitter contest station where great amounts of RF on several frequencies are present. The Radio Amateur who lives near a fellow ham operating at the 1500-watt level will find the Receiver Guard 2000 to be a great insurance policy. Field day operators can now use their own rigs without fear of losing their front ends. Those Amateurs who use listening antennas (loops, beverages, etc.) can install the unit in the coax line from the listening antennas without fear of destroying the front end of their own radio when transmitting.

SWL'ers, who as a rule use very low Q antennas, need also to protect their expensive receiver front ends. With the ever-increasing density of RF signals, SWLers who do not protect their receivers from high power RF transmitters are clearly at risk.

The Receiver Guard 2000 has less than 0.3 dB insertion loss between 1.8 and 300 MHz. The unit is attractively packaged in a black die-cast aluminum RF-tight box measuring 3.5 × 1.25 × 1.5 inches. Three models are available: Model P, the standard protection unit with RCA type phono plugs, is priced at $29.95. Model U, the standard protection unit with UHF (SO-239) fittings, is also priced at $29.95. Model CTT, the standard protection unit (Model U) with the addition of an Alpha Delta Transi-Trap™ LT Lightning Protector is available only with UHF fittings, at $49.95. (Add $4 for shipping and handling to all prices.)

For further information, contact Design Electronics Ohio, 4925 South Hamilton Road, Groveport, Ohio 34125.

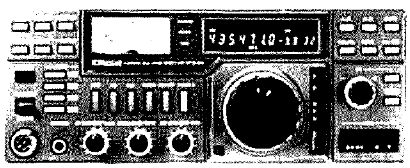
Circle #318 on Reader Service Card.

multimode transceiver

ICOM has announced the IC-471H 430-450 MHz transceiver with 75-watt transmitter and extremely low-noise PLL circuitry.

Standard features include 430-450 MHz coverage; 75 watts RF output; FM, SSB, CW modes; 32 full-function tunable memories storing frequency, offset, offset direction and tones; and 32 built-in subaudible tones, all front-panel selectable. 10 Hz tuning increments, 1 MHz up/down buttons, scanning of memories, memory modes, or band, and all mode squelch are all also standard. The compact unit features an easy-to-read fluorescent display.

The IC-471H uses 12 volt DC power and may be supplied from an external source (IC-PS15 or IC-PS30, optional) or from an optional internal AC power supply (IC-PS35). Other optional features include an IC-AG35 switchable mast-mounted preamplifier, UT15S encoder/decoder (PL encoder is standard), IC-CT10 computer interface, IC-EX309 computer interface connector and IC-EX310 voice synthesizer. The suggested retail price is $1099.



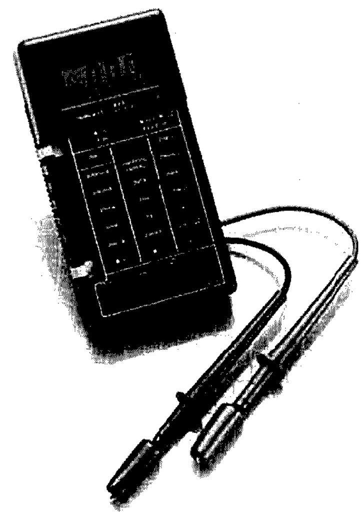
For more information, contact ICOM, 2112 116th Ave., N.E., Bellevue, Washington 98004.

Circle #319 on Reader Service Card.

digital VOM

The new Model 3550-A hand-held, push-button operated, digital VOM just introduced by Triplett Corporation offers ±0.25 percent accuracy on all DC ranges, plus 10 amp test capability and audible continuity tone. Designed for 2000 hours of battery life, the Model 3550-A is suited for in-field measurements on industrial, commercial or consumer electronic/electrical equipment.

NEW products



The tester features a new 2 amp, 600 volt fuse arrangement plus safety-designed test leads for optimum user protection.

Measuring only 3-1/2 × 6-3/4 × 1-1/2 inches and weighing only 10 ounces, the Model 3550-A has a half-inch, 3-1/2 digit LCD with polarity and low battery indication. The new 2 amp, 600-volt fuse arrangement prevents nuisance fuse blows in volt and ohm ranges. Overrange and auto-polarity are included.

Handy pushbuttons permit rapid selection of 6 tester functions. Ranges include: 0-1000 VDC in five ranges; 0-750 VAC in five ranges; 0-10 amps DC or AC current in six ranges including a 200 microamp range; 0-20 megohms in six ranges. Pushbutton selectable, hi-low power ohms and resistance diode check is included. Audible continuity is available on the 200 ohm range in the HIV mode of ohms operation.

The molded black, high-impact thermoplastic case has a sure-grip "finger tread" surface finish. An optional tilt stand facilitates bench use and easy external battery and fuse access. Other optional accessories include vinyl carrying case, battery cover, high-voltage probe, external shunt, temperature probe, clamp-on AC ammeter and line separator.

Priced at $85.00, the new Model 3550-A is warranteed for one year and is furnished complete with 9 volt (NEDA 1604) battery, 42 inch test leads, screw-on alligator clips, and a comprehensive instruction manual.

For further information, contact Triplett Corporation, One Triplett Drive, Bluffton, Ohio 45817.

Circle #320 on Reader Service Card.

flea market

RATES Noncommercial ads 10¢ per word; commercial ads 60¢ per word **both payable in advance**. No cash discounts or agency commissions allowed.

HAMFESTS Sponsored by non-profit organizations receive one free Flea Market ad (subject to our editing) on a space available basis only. Repeat insertions of hamfest ads pay the non-commercial rate.

COPY No special layout or arrangements available. Material should be typewritten or clearly printed (**not** all capitals) and must include full name and address. We reserve the right to reject unsuitable copy. **Ham Radio** cannot check each advertiser and thus cannot be held responsible for claims made. Liability for correctness of material limited to corrected ad in next available issue.

DEADLINE 15th of second preceding month.

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QSLs & RUBBER STAMPS — Top Quality! Card Samples and Stamp info — 50¢ — Ebbert Graphics 5R, Box 70, Westerville, Ohio 43081

NH ALL-BAND QTH. South slope Crotched Mountain, Greenfield. Lodge and bunkhouse, 920' elevation, 1200' from highway and power mains, 59 acres, privacy, paved access, base for future tennis court or heliport $95,000. "CW" Farr, W1WMK, Broker, Greenfield, NH. (603) 547-2053

"**KT5S**" Super DX sloper 80-10M only $59.95. "**KT5B**" Multiband dipole 160-10M only $59.95 2-KW roller inductor (28uh) $47.50. Weather boot kit (PL-259) $8.95 pp. Much more! Info available Kilo-Tec, PO Box 1001, Oakview, CA 93022. Tel: 805-646-9645.

IBM-PC RTTY ASCII/BAUDOT/CW send and receive. Split screen, buffers and features beaucoup. SASE to E. Alline, 773 Rosa, Metairie, LA 70005

WANTED: Old RCA, Western Electric, tubes, speakers, amps. (713) 728-4343, Corb, 11122 Atwell Houston, Texas 77096.

ENJOY SATELLITE TELEVISION. Save money with easy, guaranteed, do-it-yourself antenna plans/kits. Electronic knowledge not necessary. Send $1.00 for catalog or $8.95 for "Consumer Guide to Satellite Television". GFI-41, Box 9808, Missoula, MT 59807.

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THE US QSL SERVICE IS FREE. Send your QSLs to USA Hams via USQS/KM7Z, P.O. Box 814, Mulino, OR 97042. Send SASE for return QSLs and info.

TRAVEL-PAK QSL KIT — *Converts post cards, photos to* QSLs. Stamp brings circular. Samco, Box 203-c, Wynantskill, New York 12198.

AMATEUR RADIO'S NEWSPAPER — WORLDRADIO. Latest info. One year subscription (12 issues) only $10. Worldradio, 2120-B - 28th Street, Sacramento, CA 95818.

RADIO ITEMS before 1930 wanted. Buying battery operated radios, horn and cone speakers, radio tubes and parts, radio literature — books, catalogs, magazines, radio advertising signs, posters. Gary Schneider, 6848 Commonwealth Blvd., Parma Heights, Ohio 44130.

HELP! Have Model EBC-144 Jr. made by Emergency Beacon Corp. Need logics, schematics and/or maintenance manuals for this rig. Frank, WB4CIZ.

DIGITAL AUTOMATIC DISPLAYS for FT-101's TS-520's, and most others. Six 1/2" digits. Write for information. Grand Systems, P.O. Box 2171, Blaine, Washington 98230. (604) 530-4551.

RUBBER STAMPS: 3 lines $4.50 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

NOTICE: Your ads seen daily on our Computer Bulletin Board. Very low rates. Ads run for 4 weeks. 201-962-4956 to see ads. SASE for full details. Narwid BBS, 61 Bellot Road, Ringwood, NJ 07456.

"**THE SWAP LIST**" *has bargains galore. Subscribe now!* 6 months for $4.00; 1 year only $6.50. The Swap List, Box 988-H, Evergreen, CO 80439.

SCHEMATICS: Radio receivers 1920/60's. Send name brand, model, SASE. Scaramella, P.O. Box 1, Woonsocket, RI 02895-0001.

FOR SALE: Swan 350 transceiver w/ps $250. SA2040 antenna tuner $125. OF-1A audio filter $50. Kantronics CW/RTTY interface for Vic-20 with programs board + cables $135. Shipping included. Send money order. Package price $500. Write Jim Howell, KA4EBW, 18 Dan St., Salisbury, NC 28144. (704) 637-0313 evenings.

3 KW ANTENNA TUNER WM. Nye MB-1V-01 $349.00. New box unopened. Memory keyer SKM-001 $165.00. W4LNI, 3016 Cordelia, Tampa, FL 33607. (813) 876-5531.

REPAIR, ALIGNMENT, *calibration. Collins written estimates* $25; non Collins $50. K1MAN. (207) 495-2215.

INTERNATIONAL MORSE CODE TRAINER for your Commodore or unexpanded Vic Computer. Menu-driven + documentation + random tests + adjustable speed (1-25 WPM) and pitch + enters characters and hear the Morse sound. 64 version has additional features: 9 detailed lessons + user-defined tests + straight key simulator. 20 version $6.00 tape only. 64 version $14.95 tape or diskette. AC3L Software, Box 7, New Derry, PA 15671.

CHASSIS and cabinet kits. SASE K3IWK.

1985 CALLBOOKS: Prepublication orders this month. Either $16.00. Both $29.00. "Low/Medium Frequency Scrapbook", Ken Cornell, 4th edition, first printing, $7, 2/$12, dealers 10/$40. Postpaid 50. Century Prints, 6059 Essex Street, Riverside, CA 92504. (714) 687-5910.

NATIONAL RADIO CO equipment manuals price list SASE. Dust covers, NCX 3 or NCX 5 plus NCX A, pair $8.95 PP. Maximilian Fuchs, 11 Plymouth Lane, Swampscott, MA 01907.

WANTED: RTTY/CW Software for Osborne-I. Tom Yocom, 21 Bayberry Road, Acton, MA 01720.

ELECTRON TUBES: *Receiving, transmitting, microwave* ... all types available. Large stock. Next day delivery most cases. Daily Electronics, 14126 Willow Lane, Westminster, CA 92683. (714) 894-1368.

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WANTED: Cash paid for used speed radar equipment. Write or call: Brian R. Esterman, PO Box 8141, Northfield, Illinois 60093. (312) 251-8901.

SELL: 1850A Iconoscope, B.O. Filament, Radiotron 201A, brass ring base with short prongs. Raytheon 01A (1934). Heath model IM5238 AC voltmeter. Hallicrafters S38-C. G.E. table radio — M-63 mfd. by RCA. W.E. 417A, 418 and other tubes. Stan, W5TPS. (501) 636-6404.

$$$$ SUPER SAVINGS on electronics parts, components, supplies, and computer accessories. Free 40-page catalog for SASE. Get on our mailing list. BCD Electro, PO Box 830119, Richardson, TX 75083. Or call (214) 690-1102.

RCA Volt-Ohmyst WV-97A $30. WV-77C $20. Old HRO coil set $70. Beckman PH meter $25. K6KZT, 2255 Alexander, Los Osos, CA 93402.

WANTED: Old microphones, remote mixers other mic related items. All pre-1935. Box Paquette, 107 E. National Avenue, Milwaukee, WI 53204.

PORTABLE 2-meter Quads and J-Verticals. Write Radio Engineers, 3941 Mt. Brundage Avenue, San Diego, CA 92111.

RECONDITIONED TEST EQUIPMENT $1.00 for catalog. Walter, 2697 Nickel, San Pablo, CA 94806.

FOX-TANGO Newsletters — Since 1972, the prime source of modifications, improvements, and repair of Yaesu gear, free to Club members. Calendar year dues still only $8 U.S., $9 Canada, *$12 elsewhere. Includes five year cumulative index by model numbers, or send $1 for index and sample Newslet*ter. Fox Tango Club, Box 15944, W. Palm Beach, FL 33416.

RTTY-EXCLUSIVELY for the Amateur Teleprinter. One year $7.00. Beginners RTTY Handbook $8.00 includes journal index. P.O. Box RY, Cardiff, CA 92007.

IMRA International Mission Radio Assn. helps missioners — equipment loaned; weekday net, 14.280 MHz, 2-3 PM Eastern. Br. Frey, 1 Pryer Manor Rd., Larchmont, NY 10538.

"**HAMS FOR CHRIST.**" Reach other Hams with a gospel tract sure to please. Clyde Stanfield, WA6HEG, 1570 N. Albright, Upland, CA 91786.

TENNATEST — Antenna noise bridge — out-performs others, accurate, costs less, satisfaction guaranteed, $41.00. Send stamp for details, W8URR, 1025 Wildwood Road, Quincy, MI 49082.

WANTED: Early Hallicrafter "Skyriders" and "Super Skyriders" with silver panels, also "Skyrider Commercial", early transmitters such as HT-1, HT-2, HT-8, and other Hallicrafter gear, parts, accessories, manuals. Chuck Dachis, WD5EOG, The Hallicrafter Collector, 4500 Russell Drive, Austin, Texas 78745.

VERY in-ter-est-ing! Next 4 issues $2. Ham Trader "Yellow Sheets", POB356, Wheaton, IL 60189.

Coming Events

ACTIVITIES

"Places to go..."

RADIO EXPO '84 sponsored by the Chicago FM Club, Saturday and Sunday, September 22 and 23, Lake County Fairgrounds, Rt. 120 & 45, Grayslake, IL. Major manufacturers and gigantic outdoor flea market. Flea market opens 6 AM. Exhibits 9 AM. Free parking and overnight camping. Reserved indoor flea market $5/day. Tickets $3.00 advance, $4.00 at gate, good for both days. Seminars, technical talks and ladies' programs. Talk in on 146.16/76. SASE to Radio Expo '84, Box 1532, Evanston, IL 60204 or (312) 582-6923.

ONTARIO, CANADA: The Radio Society of Ontario's 16th annual Convention, October 5, 6 and 7, Westin Hotel, Ottawa. Friday night eyeball and dance. Saturday and Sunday technical sessions, demonstrations and commercial exhibits. Saturday night banquet and dance. For information: RSO Convention Committee, PO Box 15806 Station "F", Ottawa, Ontario K2C 3S7.

MARYLAND: The Columbia Amateur Radio Association's 8th annual Hamfest, Howard County Fairgrounds, Sunday, October 7, 8 AM to 3:30 PM. Admission $3.00. XYLs and children free. Reserved tables $6.00 by September 30. $8.00 after September 30. Outdoor tailgating $3.00. Indoor tailgating $6.00. Food available. Talk in on 147.735/135, 146.52/52. For tables and information: Mike Vore, W3CCV, 9098 Lambskin Lane, Columbia, MD 21045. 992-4953.

INDIANA: The 5th annual Grant County ARC Hamfest, Saturday, September 8, McCarthy Hall, St. Paul's Catholic Church, Marion. Doors open 8 AM. Donation $2.00 advance, $3.00 gate. Refreshments, free parking. 8 ft. tables $2.00. Talk in on 146.19/79 and 146.52 simplex. For information/tickets SASE to: WD9EOI, Jim Allman, 1108 Spencer Avenue, Marion, IN 46952.

PENNSYLVANIA: The Mt. Airy VHF ARC (Pack Rats) invites all Amateurs and friends to the 8th annual Mid-Atlantic VHF Conference, Saturday, October 6, Warrington Motor Lodge, Rt. 611, Warrington. And the 13th Pack Rat Hamarama, Sunday, October 7, Bucks County Drive-in Theater, Rt. 611, Warrington. Flea Market admission $3.00. Selling spaces $5.00 each. Gates open 6 AM. Rain or Shine. Bring your own tables.

Advance registration for the Conference including Hamarama admission $4.00. Send to Hamarama '84, P.O. Box 311, Southampton, PA 18966 or Lee A. Cohen, K3MXM (215) 635-4942.

NEW YORK: The Elmira Amateur Radio Association's 9th annual International Hamfest, Saturday, September 29, Chemung County Fairgrounds. Gates open at 6 AM until 5 PM. Outdoor flea market. Indoor dealer displays of new equipment. Breakfast and lunch available on premises. Tickets available at the gate or in advance from Steve Zolkosky, 118 East 8th St., Elmira Heights, NY 14903.

GEORGIA: The 11th annual Lanierland ARC Hamfest, Sunday, September 23, 9 AM in the Holiday Hall of Holiday Inn, Gainesville. Flea market, left foot CW contest and many other activities. Free tables and inside display area for dealers reserving in advance. Doors open 8 AM for setup. Talk in on 146.07/.67. For information: Phil Loveless, KC4UC, 3574 Thompson Bend, Gainesville, GA 30506. (404) 532-9160.

TEXAS: Tornado Alley Hamfest, sponsored by the Wichita Amateur Radio Society, September 22 and 23, National Guard Armory, Wichita Falls. Saturday 9 to 5; Sunday 9 to 2. Dealer displays and demonstrations. Large inside flea market. Ladies' activities. Nearby museums, art center and shopping. Pre-registration $4.00. $5.00 at the door. Air Force MARS, QCWA meeting, QLF contest, homebrew contest and more. Talk in on 146.34/94, 147.75/15, 449.30/444.30 and 449.20/444.20. For information: Wichita Amateur Radio Society, P.O. Box 4363, Wichita Falls, TX 76308.

GEORGIA: The Amateur Radio Club of Augusta'a annual Hamfest, Sunday, September 16, Julian Smith Casino Park. Refreshments, Bar-B-Q, entertainment. Dealers welcome. Flea Market Tickets $1.00; 6/$5.00; 13/$10.00. Talk in on 145.49-600. Hospitality room Saturday evening, Ramada Inn West, Washington Road, Rm 108-110. For information SASE to: D.F. Miller, 4505 Shawnee Rd., Martinez, GA 30907. (404) 860-3700.

PENNSYLVANIA: The Skyview Radio Society's annual Hamfest, Sunday, September 16, noon to 4 PM, Club Grounds, Turkey Ridge Road, New Kensington. Registration $2.00. Vendors $4.00.

NEW MEXICO: Northern New Mexico Hamfest, Sunday, October 7, 8 AM to 3 PM, Terrero Group Shelter along the Pecos River east of Santa Fe. Tailgate flea market, meetings, fishing, picnicking, family fun. Admission $3. Children $1.50 includes hot dogs, chips and free Saturday night camping. Talk in on local repeaters and 52 simplex. For information SASE to: Northern New Mexico ARC, c/o Bob, N5EPA, Rt. 3, Box 95-15, Santa Fe, NM 87501 or call on 3.939 MHz at 0100 UTC.

ILLINOIS: The Peoria Area Amateur Radio Club's Superfest '84, September 15 and 16, Exposition Gardens, W. Northmoore Rd., Peoria. Gates open 6 AM, Commercial Building at 9 AM. Admission $3.00 advance, $4.00 gate. Children under 12 free. Amateur Radio and computer displays, huge flea market, free bus to Northwoods Mall on Sunday. Full camping facilities on grounds. Saturday night informal get-together at Heritage House Smorgasboard, 8209 N. Mt. Hawley Rd., Peoria. Talk in W9UVI on 146.16/76. For information and reservations SASE to Superfest '84, P.O. Box 3461, Peoria, IL 61614.

NEW YORK: Electronics Fair and Giant Flea Market, sponsored by the Yonkers ARC, Sunday, October 7. 9 AM to 4 PM, rain or shine. Yonkers Municipal Parking Garage, Corner of Nepperhan Avenue and New Main Street. All day demonstrations; Amateur Radio, computers, electric car, satellite TV, SSTV and more. Giant auction 2 PM. Refreshments, free parking, facilities. Free coffee all day. Admission $2.00. Children under 12 free. Sellers $6.00 per space admits one. Bring tables. For information: YARC, 53 Hayward Street, Yonkers, NY 10704. (914) 969-1053. Talk in on 146.265T-146865-R or 52 direct. CB channel 4.

VIRGINIA: ARRL Roanoke Division Convention and 9th annual Amateur Radio-Computer Fair, Saturday and Sunday, September 22 and 23, Virginia Beach Pavilion. 9 AM to 5 PM. Displays, forums, computer equipment, giant flea market, ladies' activities, movies for the kids. Admission $4.00 advance (good for both days). $5.00 at door. Flea market tables $5.00 one day, $8.00 both days. Plan a family vacation at beautiful Virginia Beach. Visit the Waterside Festival Marketplace in Norfolk with its specialty shops and restaurants. For information/tickets: Jim Harrison, N4NV, 1234 Little Bay Avenue, Norfolk, VA 23503. (804) 587-1695.

CONNECTICUT: The Natchaug Amateur Radio Association's annual Giant Flea Market. September 23, Elks Home, 198 Pleasant Street, Willimantic. Starts 9 AM. Dealers 8 AM. Admission $2.00. Under 16 free. Advanced reserved tables $5.00. $7.00 at door. Plenty of food and drink. Talk in on 52 direct and 147.30/90 repeater. For information: Ed Sadeski, KA1HR, 49 Circle Drive, Willimantic, CT 06226. (203) 456-7029 after 4 PM.

NEW HAMPSHIRE: The Connecticut Valley FM Association's 8th annual Hamfest and Flea Market, September 16, King Ridge Ski Area, Sutton. 9 AM to 5 PM, rain or shine. Admission $2.00. Dealers and flea market $3.00 per tailgate or table. Food available on premises. Overnight camping for self-contained units only. No hookups. Talk in on 146.16/76 or 146.52 simplex.

KENTUCKY: The 14th annual Greater Louisville Hamfest and Great Lakes Division Convention, Saturday and Sunday, September 29 and 30, Kentucky Fair and Exposition Center. 8 AM to 5 PM both days. Air-conditioned indoor exhibitors' area and flea market. Meetings and forums. Hotels across from Hamfest site. Camping available on grounds. For information: Greater Louisville Hamfest Association, P.O. Box 34444, Louisville, KY 40232. (502) 368-6657.

ALABAMA: Hospitality Hamfest sponsored by the Mobile ARC, September 15 and 16, Texas Street Recreation Center off I-10, Mobile. Doors open 9 AM. Admission free. Activities for ladies, swap tables, parking, good food and fellowship. Talk in on 146.22/82. For information: Porter Chambers, KI4FE, 3320 Emelye Drive, Mobile, AL 36609. Call 661-1160.

TENNESSEE: The fourth annual Tri-Cities Hamfest, sponsored by the Johnson City, Kingsport and Bristol Amateur Radio Clubs, Saturday, October 20, Appalachian Fairgrounds, Gray. Forums, dealers, flea market and RV hookups. For information: Tri-Cities Hamfest, PO Box 3648 CRS, Johnson City, TN 37601.

NEW YORK: Ham-O-Rama and Computerfest '84, Friday evening, September 7, 6 PM to 9 PM and Saturday, September 8, 7 AM to 5 PM, Erie County Fairgrounds, Buffalo Raceway, south of Buffalo. Indoor/outdoor flea markets, new equipment and video displays, computer demonstrations, tech and non-tech programs. Chicken barbeque, awards and more. Admission $3.50 advance. After August 24 and at gate $4.50. Outside flea market $3.00; inside $10.00. Talk in via W2EUP/R 146.31/91 and 146.52. For information: Nelson Oldfield, 126 Greenway Blvd., Cheektowaga, NY 14225.

CALIFORNIA: The Sonoma County Radio Amateur's second annual Ham Radio Flea Market, Saturday, September 15, 8 AM to 2 PM, Sebastopol Community Center, 390 Morris St., Sebastopol, 5 miles west of Santa Rosa. Admission and parking free. Tables $6 at door or $5 advance. Vendor set up 7 AM. Radio clinic, exhibits, refreshments, auction at noon. Talk in on 146.13/73. For tickets/information: SCRA, Box 116, Santa Rosa, CA 95404.

TENNESSEE: Memphis Hamfest, sponsored by the Mid South Amateur Radio Association, Delta Radio Club and Memphis Radio Relay Club, October 13 and 14, Pipkin Building, Memphis Fairgrounds. 8 AM to 4 PM Saturday, 9 AM to 2 PM Sunday. Forums, ladies' activities and large flea market all inside in air-conditioned comfort. Flea market tables $5.00 each per day. Trailer hookups available. For information: Clayton Elam, K4FZJ, 28 No. Cooper, Memphis, TN 38104 (901) 274-4418 Days. (901) 743-6714 Nights.

NEBRASKA: 8th annual 3900 Club Hamboree and Iowa State Convention, October 12 and 13, Marina Inn, South Sioux City. Sponsored by the 3900 Club and Siouxland Amateur Radio Repeater Association. Flea market, exhibits, ladies' programs. Air Force MARS, QCWA, UHF/VHF, ARRL, DX session. Novice session and QSL Bureau. Friday night get-together. Saturday night banquet, Dr. Beverly Mead, speaker. Flea market and convention $6.00. Banquet $10.00 ($12.00 at the door). Flea market tables $4.00 each. All indoor facility. Talk in on 146.37/146.97. For advance reservations: Dick Pitner, 2931 Pierce, Sioux City, IA. Advance flea market reservations: Al Smith, 3529 Douglas, Sioux City, IA.

MASSACHUSETTS: The 1979 Amateur Radio Association is sponsoring Novice and Technician/General classes starting September 18 at the Chelsea High School, Chelsea, MA. Admittance is free. Student pays cost of materials. For more information: Frank, K1BPN, 1979 ARA, P.O. Box 171, Chelsea MA 02150.

NEW ENGLAND: Hosstraders' Fall Tailgate Swapfest, Saturday, October 6, sunrise to sunset at Deerfield, NH Fair grounds. Admission $2 including tailgaters. Friday night camping at nominal fee after 4 PM. No reservations. Profits benefit Boston Burns Unit of Shriners Hospital. Last Spring's donation $5,813.00. For map to northeast's biggest ham flea market SASE to Norm, WA1IVB, RFD Box 57, West Baldwin, ME 04091.

OPERATING EVENTS

"Things to do..."

SEPTEMBER 8: The Mark Twain ARA will operate W0KEM from 1400Z to 2300Z, Sept. 8 and 9 to celebrate the dedication of the 20,000 acre Mark Twain Lake and Clarence Canon Dam in East Central Missouri. Phone: lower 25 kHz of 40, 20 and 15 meter General band. Novice operation in 40 meter band. For a certificate send legal SASE to Mark Twain ARA, P.O. Box 56, Center, MO 63436.

SEPTEMBER 8 AND 9: "WE TALK SO THEY CAN WALK." The Ararat Shrine Radio Club of Kansas City, MO, will hold a talk in to benefit the crippled children's hospitals. We will host a multi-band, multi-operator talk in from 10 AM to 6 PM each day. First 10 kcs of general portion of ham bands and first 10 kcs of 40 meter Novice band. For any contact with club station, WA0NQA, you will receive a two-color certificate with name and call. Send $1.00, QSL card and large SASE to: J.V. Foust, KA0GBK, 5240 N. Palmer, Kansas City, MO 64119. All monies will go to the crippled children's hospitals. Your QSL card will be displayed in the Kansas City Shrine Temple Radio room.

SEPTEMBER 8 AND 9: The Radio Association of Erie (W3GV) will commemorate Admiral Perry's victory at the Battle of Lake Erie during the War of 1812. 1200Z to 0100Z Saturday and 1200Z to 2100Z Sunday. 7.235, 14.235 MHz (phone) and 7.090, 14.090 MHz (CW/RTTY). Special QSL and historical date on the flagship Niagara via W3GV, 4572 Southern Dr., Erie, PA 16506 or W3 QSL Bureau for DX stations. Please enclose business SASE.

SEPTEMBER 15: The McHenry County Wireless Association will sponsor the 2nd annual DXpedition to Cedar Island, Fox Lake, Illinois. Operation begins at 10000 CDT on lower 20 kHz of phone portion of 40 and 15 meters. An attractive QSL card will be supplied for all confirmed contacts.

SEPTEMBER 22: The Paul Bunyan Wireless Association and the Brainerd Area Amateur Radio Club will sponsor a special event station from the site of the Paul Bunyan Festival near Brainerd, MN, from 1800Z on September 22 to 2100Z on September 23. Lower portion of the General class phone portion of 40-10 meters. Send QSL and SASE to KC0YG for a commemorative QSL.

SEPTEMBER 22: The Alford Memorial Radio Club of Stone Mountain, Georgia, will sponsor its first annual Pig Out from 0400 to 2200Z. SSB phone and CW, 10 kHz above bottom of General portion of 80-10 meters. For a commemorative QSL and special certificate for contact, SASE with information to: Alford Memorial Radio Club, P.O. Box 1282, Stone Mountain, GA 30086.

OCTOBER 13 AND 14: Columbus Day International DX Contest in commemoration of Columbus Day, sponsored by the Miami Havana Lions Club. From 1200 GMT Saturday to 2400 GMT Sunday. Any Amateur station making five contacts with official Radio Club DX member operator during the 2 days will be eligible to apply for the Miami Havana Lions Club OSL award. Exchange RS(T) and QTH. For this special award send QSL's or log and $2.00 U.S. funds or 6 IRC's to: Miami Havana Lions Club, Box 674, Miami, FL 33135. At the start of the contest, October 13, 1200 GMT, members of the Contest Committee will read the names and assigned numbers of the official operators in the following frequencies: 28.915, 21.250, 14.250, 7.230 phone.

A FAR NET AWARD CERTIFICATE offered by the Armored Force Amateur Radio Net. Non-member stations qualification requirements: For basic award, non-member stations must establish 2-way contact with a minimum of 15 different A FAR NET member stations. For endorsements, non-member stations must contact ten or thirty-five additional members, any band, any mode. Confirmation of required contacts through copy of log certified by two other Amateur radio operators. Send application with 50¢ minimum for postage, etc. to: Alfred G. Beutler, K2DWI, A FAR NET Certificate Manager, 36 Manchester Road, East Aurora, NY 14052.

SPACE SHUTTLE COMMENTARY VIA OSCAR SATELLITE. The Spaceport Amateur Repeater Club (SPARC) has been authorized by AMSAT to transmit Space Shuttle mission commentary for all missions on Special Services channel H2. 145.963 MHz of AMSAT-OSCAR 10. All Amateur Radio operators are invited to submit reception reports to: SPARC, P.O. Box 672, Merritt Island, FL 32952.

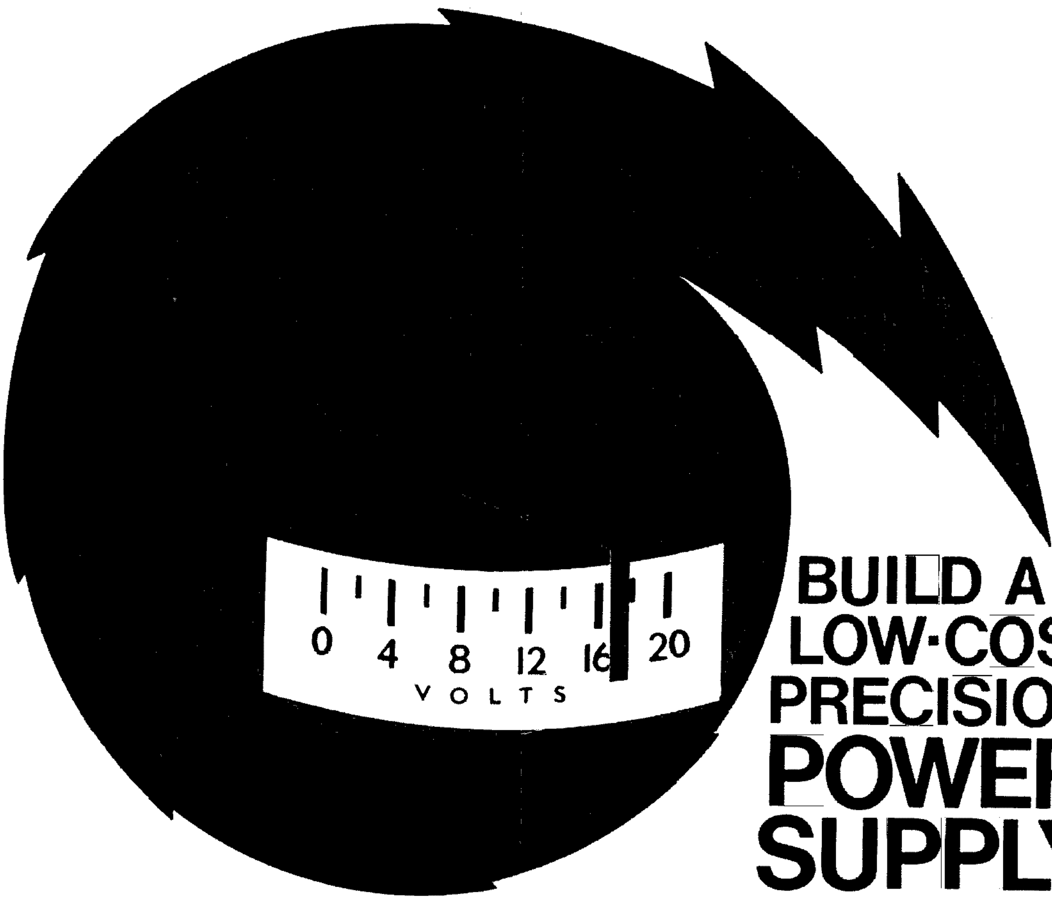
AUGUST 31 AND SEPTEMBER 1: The Wireless Institute of New Orleans (W.I.N.O.) will operate K5WF from the Louisiana World Exposition — World's Fair. 10 AM to 10 PM CDT daily on HF bands, all modes and 40 meters. LSB about 7.240 MHz. Also 75 and 20 meters, propagation allowing. A special commemorative QSL/Certificate confirming contacts will be available for a SASE: W.I.N.O., Box 6541, New Orleans, LA 70174.

FRONT COVER: 70 Years of SSB

An early single sideband receiver, designed by James Lamb in 1937, included provisions for duplex operation on 4 MHz and used the filter method of SSB generation. Shown within the schematic for Lamb's receiver is a state-of-the-art SSB receiver. Developed by RCA Corporation Government Communications Systems under the guidance of Dr. Ulrich L. Rohde, business area director of radio systems, the unit is a highly software-based experimental transceiver with frequency-hopping capabilities. It contains a special receiving system for noise blanking and uses a delay line to enable processing. The transceiver covers 1.5 to 30 MHz in transmit and 10 kHz to 30 MHz in receive in 100 Hz steps. Modes of operation include AM, CW, SSB, FSK, and frequency hopping. Particular software features are automatic FFT (Fast Fourier Transform) analysis of signals and tuning of SSB with an error of less than 1 Hz, frequency and memory scanning, real-time clock operation, and storage of 100 frequencies with all signal parameters. It utilizes a second alphanumeric display with a self-help feature that tells the user how to activate and use the software-based function. It has an intercept point of greater than +30 dBm and an SSB phase noise of −140 dB at 25 kHz to −160 dB at 100 kHz, indicating dynamic performance. AGC is 0.2 microvolts to 1 volt, with a 4 dB audio change and linearized S-meter look-up table in ROM.

(These are just a few of the outstanding features/controls available on the front panel. Ed.)

ham radio magazine



BUILD A LOW-COST PRECISION POWER SUPPLY

focus on communications technology

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contents

presstop de W9JUV

A PROPOSAL TO REALLOCATE 220 MHZ AMATEUR FREQUENCIES TO LAND MOBILE was made in late July by Sideband Technology, Inc., a manufacturer of Amplitude Compandored Sideband (ACSB) equipment. In its Petition for Rule Making, RM-4831, Sideband Technology proposes that the bottom two megacycles of the 220-225 MHz Amateur band plus the adjacent 216-220 MHz Inland Waterways Communications Service band be allocated to ACSB Land Mobile users.

Opposition To The Growing Pressure On 220 Has Developed Rapidly, highlighted by a teleconferencing net August 2 that carried the story throughout the country via many VHF and UHF repeaters, plus a few SSB relays to the HF bands. The net, hastily organized and led by 220 Notes publisher K9XI, did an excellent job of reviewing the growing 220 problem and what could be done about it. It was particularly valuable in alerting Amateurs to file Comments on both this petition and the earlier one by the Land Mobile Communications Council (see September Presstop). Comments closed on the LMCC petition in August, but a legal question has delayed the closing date on RM-4831 indefinitely.

A Reallocation Of This Importance Is Highly Unlikely without going through a formal Notice of Proposed Rule Making procedure, but rumors of high-level FCC support for the change have been circulating. (See this month's Reflections for more on the 220 issue).

AN ELECTION OF UNIONIZING ARRL EMPLOYEES WAS HELD August 23 in Newington, and the union effort lost by a very large margin. Staff dissatisfaction over ARRL's return to the five-day work week and other work-related matters had triggered the effort to set up a collective bargaining agent at League headquarters, but a very strong, well-organized opposition effort by management defeated the attempt. However, the problems that caused the unionizing effort have still not been resolved. That, coupled with the deterioration in relations between staff and management—and within those groups as well—that developed during the unionizing efforts of the past few months make it likely that morale will continue to be a problem in Newington for some time to come.

420-450 MHZ USERS IN GEORGIA AND TEXAS FACE POWER RESTRICTIONS following the addition of two new "Military Protected Zones." After September 12, Amateurs within a 124 mile radius of Warner Robbins AFB in Georgia or Goodfellow AFB in Texas who plan to run more than 50 watts ERP will be required to coordinate their operations on the 70 cm band with both the nearest FCC Engineer-In-Charge and the military Area Frequency Coordinator.

OSCAR 10'S OPERATING SCHEDULE HAS BEEN CHANGED, due primarily to the onset of an eclipse period which deprives the spacecraft solar panels of sunlight for up to 1 1/2 hours at a time. The 145.810 General Beacon schedule has also been changed; bulletins now start on the hour and half hour with five minutes of CW, followed by 10 minutes of PSK telemetry, then RTTY bulletins (50 baud) for another five minutes. PSK telemetry occupies the final 10 minutes of each half-hour sequence. For schedule updates check the AMSAT nets.

The 75-Meter AMSAT Net Frequencies Have Been Changed to permit General licensees to take part in the Tuesday night nets following the September 1 phone band expansion. At first the 0100Z (Wednesday, GMT) nets moved to 3855 kHz, but problems with an existing operation on that frequency caused a further move. Look for them around 3856-3860 kHz.

ACSB Experiments Through OSCAR 10 began August 24 in a cooperative AMSAT, ARRL, and Project OSCAR effort. Expected benefits include improved signal-to-noise ratios (narrower bandwidth) and easier tuning since ACSB receivers lock onto a pilot carrier and thus automatically track doppler shift. ACSB can be received with a normal SSB receiver, despite the 3.1 kHz pilot carrier. Sideband Technology is supplying the ACSB equipment.

AMATEUR ACCESS TO THE NEW 24 MHZ WARC BAND MAY NOT BE TOO FAR OFF. There had even been some thought that the FCC would begin 24 MHz implementation before the Commissioners went on August recess, as there seems to be no problem with current band occupants. On the other hand, the 18 MHz band still supports considerable commercial activity so is not likely to be reallocated very soon.

902-928 MHz's Availability Is Also Likely in the near future, and the ARRL VHF-UHF Advisory Committee has a band plan worked up for it. However, before either 18 or 902 MHz becomes available specifics such as modes and subbands must be established through an NPRM.

COMBINING METEOR SCATTER WITH PACKET RADIO could prove a viable mode for Amateur communications, suggests ARRL's QEX. WØRPK is on 50.505 MHz from central Iowa, and K1HTV planned some 145.05 MHz tests during the August Perseids shower. W1AW may also gear up for both 10 and 6 meter meteor scatter packet radio experiments in the near future.

ARRL IS SUPPLYING ANSWERS TO ALL AMATEUR EXAM QUESTIONS in an attempt to establish uniform exams throughout the country. They distributed Element 3 (Tech/General) questions complete with distractors to other VECs and Amateur Radio publishers in August, and expect to have the Advanced and Extra sets out soon. Whether those VECs who've already invested considerable time and effort in working up their own answer/distractor sets will be willing to change at this time remains to be seen, however.

REFLECTIONS

the endangered spectrum: 220 MHz under fire

I. Who's on 220?

The 1-1/4 meter Amateur band — covering 220-225 MHz — is in imminent danger of being lost to commerical interests.

So what? Who operates there anyway?

This is the substance of the argument presented by Sideband Technology, Incorporated, in its petition to the FCC (RM-4831). STI, who manufactures amplitude compandored sideband (ACSB) equipment for the land mobile industry, is suggesting that because hams hardly use 220-222 MHz, that segment of the band should be reserved for land mobile purposes instead. A second group, the Land Mobile Communications Council (LMCC) has also petitioned the FCC (in RM-4829), suggesting that the *entire* band should be exclusively reserved for land mobile use.

"T'aint so!" For all practical purposes, the band is divided into two parts. 220-222 MHz is used by Radio Amateurs doing pioneering radio work (EME, propagation beacons, weak-signal communications, packet radio, and experimental and control links, etc. 222-225 MHz is used for FM repeater and general CW/SSB operation.

Just how did the commercial interests get the impression that the lower portion of the band — where the bulk of Amateur experimentation is being carried out — was unoccupied? *They conducted an exhaustive study by consulting a* ***repeater directory.***

What can *you* do?

1. Take this threat SERIOUSLY.
2. Read the additional information provided by W9JUV and W1JR on these pages.
3. Weigh the facts — check to see what activity exists in your area — and then . . . *be ready to respond* ***rapidly*** *in writing to the FCC if it gets to the NPRM (Notice of Proposed Rule Making) stage.**

Though the formal comment period for one of the petitions ended as this issue was going to press August 29, we believe the FCC would still find your comments valuable. If you do respond, be sure to indicate that you wish them to accept this as a *late comment*, and that your comment is ***in opposition*** to a specific petition (either RM-4829 or RM-4831). State your name, call sign, class of license, and the year you were first licensed. Provide as much information about 220 MHz activity in your area — repeaters, control links, weak-signal work, etc. — as possible. Type (double-space) and stick to the facts.

You will have an opportunity to file a formal comment when and if an NPRM on the reallocation of the 220-MHz band is issued by the FCC. If this occurs, a well documented, logical argument will carry the greatest weight.

Ham radio will keep you informed of late-breaking events, as will other news services; follow *Westlink Report* (via mail or repeater), the *W5YI Report*, *ARRL letter*, and ARRL bulletins. Do your homework, be prepared, and respond formally at the appropriate time.

Rich Rosen, K2RR
Editor-in-Chief

*A comprehensive report on all aspects of this situation, including those parts of the actual petitions pertaining to land mobile interest in the 220 band is available in the latest issue of *220 Notes*. (308 Eastgate Court, New Lenox, Illinois 60451 — single copy, $1.25; yearly subscription, $5.)

II. How We Got There — A Brief History

The 1-1/4 meter band has survived a number of changes since its establishment just before World War II. Before the war, everything above 110 MHz was designated "Amateur and experimental." During the war, the spectrum from about 200 MHz up was used for the primitive radars of that period, radio altimeters, and radio navigation aids. Shortly afterwards, the U.S. phased out most of its use of the frequencies just above 220 MHz; Amateurs were temporarily allocated a band at 235 to 240 MHz and then moved to 220 to 225 MHz in the late 1940's. In much of the world, however, military use of the band continued, and Amateurs outside of the Americas never did get a post-war 1-1/4 meter band.

Because the U.S. wanted to keep 220 to 225 — adjacent to its 225 to 400 MHz military aircraft frequencies — available for expansion, it permitted Amateur use of the band on a secondary basis. This became the standard for ITU Region 2, while the rest of the world continued to use the band for radiolocation and/or land mobile. During the 1960's and 1970's other services cast frequency-hungry eyes on 220 to 225 MHz, but were discouraged by the government's continued intention to hold on to the band, with Amateurs acting in a "caretaker" role.

WARC 79 wrought a change with far-reaching implications for Region 2 Amateurs. The U.S. government relinquished interest in the band, and the Conference reallocated these frequencies to Amateurs and fixed mobile services on a co-equal basis. Thus the die was cast.

In an FCC study conducted several years ago, 220 to 225 MHz was mentioned in passing as a possible source of land mobile frequencies. This past June, the band was again cited by the LMCC in a frequency survey submitted to the Commission. However, the first direct attack on continued Amateur use of the band came in July when S.T.I. filed its petition.

To the horror of both FM equipment manufacturers and FM land mobile users, the FCC has proposed putting ACSB channels *between* existing land mobile FM channels. Indeed, some 150 MHz equipment has been licensed by the FCC on a temporary basis and is already in operation: ACSB promoters, however, would prefer to see it occupy a place where it could be exercised without interference to or from other communications modes. 220 to 225 MHz seems to be the answer to their prayers.

Can this threat be stopped before it reaches the status of a Notice of Proposed Rule Making? Perhaps not. By rejecting the "no-code" license, we lost our chance to ensure that 220 to 225 MHz would be "off limits for commercial encroachment for years to come; most of the no-code supporters had recommended that at least part of the 220 to 225 MHz range be allocated to no-code license holders. A few voices in the wilderness even pointed out the value of new licensees to frequency preservation of that band. Had no-code been adopted, it is inconceivable that the Commission would have turned its back on the new class of Amateur license it had just created by giving that class's prime frequencies to another service. With that opportunity gone, the industry's conception of the 220 to 225 MHz frequency range is that it is under-utilized, with significant activity only in major urban areas and then on only a few repeater pairs. *"Use it or lose it"* has never been more true than it is today.

— Joe Schroeder, W9JUV

two petitions before the FCC threaten yet another Amateur band

III. Why Amateur Radio Needs 220 MHz

Despite the lack of commercially available equipment and worldwide participation, 220 has seen continued growth over the years. This growth has taken two directions — weak signal operation in the lower segment and FM operation in the upper portions of the band. The latter is understandable in terms of severe crowding in the 2 meter and 70 cm (440-450 MHz) bands — and because foliage attenuation decreases the lower you go in frequency, 135 cm (220 MHz) is preferred to 70 cm in suburban areas.

The 220 band plan written by the ARRL VUAC (VHF/UHF Advisory Committee) in 1978 (**table 1**) designated 222-225 MHz for the use of FM repeaters. This has been widely accepted. 220-222 MHz was divided up into two portions: weak-signal on 220-220.5 and FM repeater links, remote bases and control links on 220.5-222 MHz. Information on the latter stations is rarely published in order to prevent unauthorized control of the repeaters. (If STI and the LMCC had done their homework, they would have been aware of this. — Ed.)

The lower portion has been extensively used by weak-signal operators for the exploration of new propagation modes. Despite the band's claimed similarities to 2 meters, there are important differences between the two. For example, in meteor burst communication cutoff frequency is often just below or slightly above the 220 band. (It wasn't until 1968 that a 220 meteor scatter QSO was completed.) To this day, weak-signal operators are experimenting to find the cutoff frequency for the slower speed meteor showers. *Further research is essential.* For EME communication, 220 is a practical frequency because the antenna size required and the sky noise present is just about 50 percent of that at 2 meters. Auroral propagation, seldom seen at the higher frequencies such as 70 cm, is often possible.*

It was the 220 band that accounted for the 2500-mile DX contact between California (W6NLZ) and Hawaii (KH6UK) in the late 1950's via a tropo-ducting mode. More recently, in 1983, the trans-equatorial path was finally conquered when KP4EOR worked LU7DJZ over a 3670 mile (5906 km) path. Still to be explored are the FAI (Field Aligned Irregularities) and mid-latitude sporadic-E modes, which while theoretically possible have not yet been successfully used to establish two-way contact. This unique frequency range should be left intact for Radio Amateurs to further the state-of-the-art in radio wave propagation.

(Why wasn't all this activity obvious to those conducting a study on band occupancy? Perhaps the survey was performed during the regular work-week, while weak signal operators were busy working at their regular jobs, and not on the air. Also, weak-signal operators employ highly directional antennas, and would not be detected unless those conducting the survey were directly between the sending and receiving antennas.)

Nationwide activity on this band is apparent by the achievement of the WAS (50 states) award last year by a number of Amateurs. It's also easy to check the number of entries in the various VHF/UHF contests, 220 standings of active stations listed in *QST's* "The World Above 50 MHz" or correspondence in the dedicated newsletter: *220 Notes*. That weak-signal interest and activity exists is obvious, even though it may be most evident during special propagation openings, contests, and net nights. Haphazard monitoring of the band results in inaccurate data. To obtain a valid indication of 220 activity, monitoring would have to be done nationwide, and over a longer period of time; one would assume the land-mobile industry would have done so before making rash (and incorrect) statements claiming that the lower portion of the 220 band is not well used.

Only a year ago Amateurs received FCC permission to experiment with Packet Radio on the VHF FM bands. The overcrowding of 2 meters and the requirements for additional bandwidth to support 9600 baud and higher data rates make 220 a natural. (Part 97 of the FCC regulations prohibits high baud rates below 220.5 MHz. — Ed.) In the New England area, several stations using Packet Radio in the 221-222 MHz region have just become active.

This frequency range is not without problems. Amateurs in the vicinity of TV channel 13 experience video beats across the weak-signal portion of the 220 band. This is further complicated by radiation from TV set local oscillators, which often radiate into the lower portion of the 220 band when tuned to TV channel 7. No *fixed* commercial service would care to operate in the vicinity of these interfering sources. (Amateurs, on the other hand, are frequency-agile and can attempt to minimize these effects. In addition, radiation from the second harmonic of the local oscillator of the ubiquitous frequency scanners can fall within the 220 band, affecting any and all services. — Ed.)

Where can the land-mobile interests go? Considering how outdated the present FCC allocation of TV channels is, a better idea might be to go *entirely* to the UHF spectrum and make the VHF spectrum (land mobile allocated frequencies) available for point-to-point and mobile-to-base station use. The FCC could probably allocate more spectrum by taking advantage of the unused TV channels in the UHF spectrum, eliminating all channels below channel 25. (The LMCC petition, in fact,

(continued on page 119)

table 1. Proposed 220 to 225 MHz band plan. (Submitted by Joe Reisert, W1JR, Chairman VUAC.)

frequency (MHz)	allocation
220.0-220.05	EME (Earth-Moon-Earth)
220.05-220.06	propagation beacons
220.06-220.1	weak signal CW
220.1	calling frequency (Note 1)
220.1-220.5	general weak-signal, rag chewing and experimental communications
220.5-221.9	experimental and control links
221.9-222.0	weak-signal guard band (Note 2)
222.0-222.05	EME (Note 2)
222.05-222.06	propagation beacons (Note 2)
222.06-222.1	weak-signal on CW (Note 2)
222.1	calling frequency (Notes 1 and 2)
222.1-222.3	general operation CW or SSB etc. (Note 2)
222.3-222.38	FM repeater inputs (Note 3)
222.42-223.9	FM simplex (Note 4)
223.94-225.0	FM repeater outputs (Note 3)

Notes:
1. After establishing contact QSY at least 10 kHz up or down for general rag chewing or weak-signal operation.
2. This area is primarily needed where TV channel 13 spillover and long-distance search radars are in operation (e.g., California area).
3. This is the existing ARRL Repeater Plan, 1.6 MHz splits.
4. 223.5 is the national calling frequency for FM simplex. Alternative frequencies in 20 kHz increments in this region.

a six-output power supply

Build a low-cost lab-quality voltage source

This power supply construction project will provide a valuable addition to the workbench of most Amateurs. The unit includes three positive and three corresponding negative outputs consisting of a dual-tracking pair, a positive and negative independently adjustable pair, and a positive and negative programmable voltage standard pair. A detailed set of drawings for construction is provided. The total cost of completing the project is surprisingly low, mainly because the unit is built around nine op amps that cost approximately fifty cents each.

independently adjustable supply

Any voltage regulator or voltage regulated power supply must have five parts: a voltage reference, an error amplifier, a feedback mechanism, a series-pass transistor or means of current-buffering and, usually, an optional current-limiting or self-protection circuit. This power supply includes all five of these elements. In circuit B (CKTB) of fig. 1, diode D5,* a 1N754 6.8-volt zener, provides a 6.8-volt input to the noninverting (+) input of op amp U1. The (−) or inverting input will consequently also have +6.8 volts on it through normal op amp action (feedback). The voltage reference provided by op amp U1 is impressed across pot R6, a 10K pot. This variable voltage of from 0 to +6.8 volts is again applied to the noninverting (+) input of op amp U2, a 741 or LM101 type bipolar amplifier. Gain is determined by:

$$E_{out} = E_{in}\frac{(R10 + R11) + R12}{R12}$$

$E_{in} = 6.8$ volt reference voltage and is derived by pot R6 from 0 to 6.8 volts.

If you work through the numbers, you will derive an output of from approximately 0 to 16.5 volts with output adjust trimmer pot, R11, set halfway at 2500 ohms.

The op amps themselves have internal current

*Note that "D", rather than *ham radio's* customary "CR", is used to designate diodes in this article and accompanying artwork.

By Vaughn D. Martin, 114 Lost Meadows, Cibolo, Texas 78108, and **Dean Davis, WB5ZKU**, 6206 Ridge Oak, San Antonio, Texas 78250

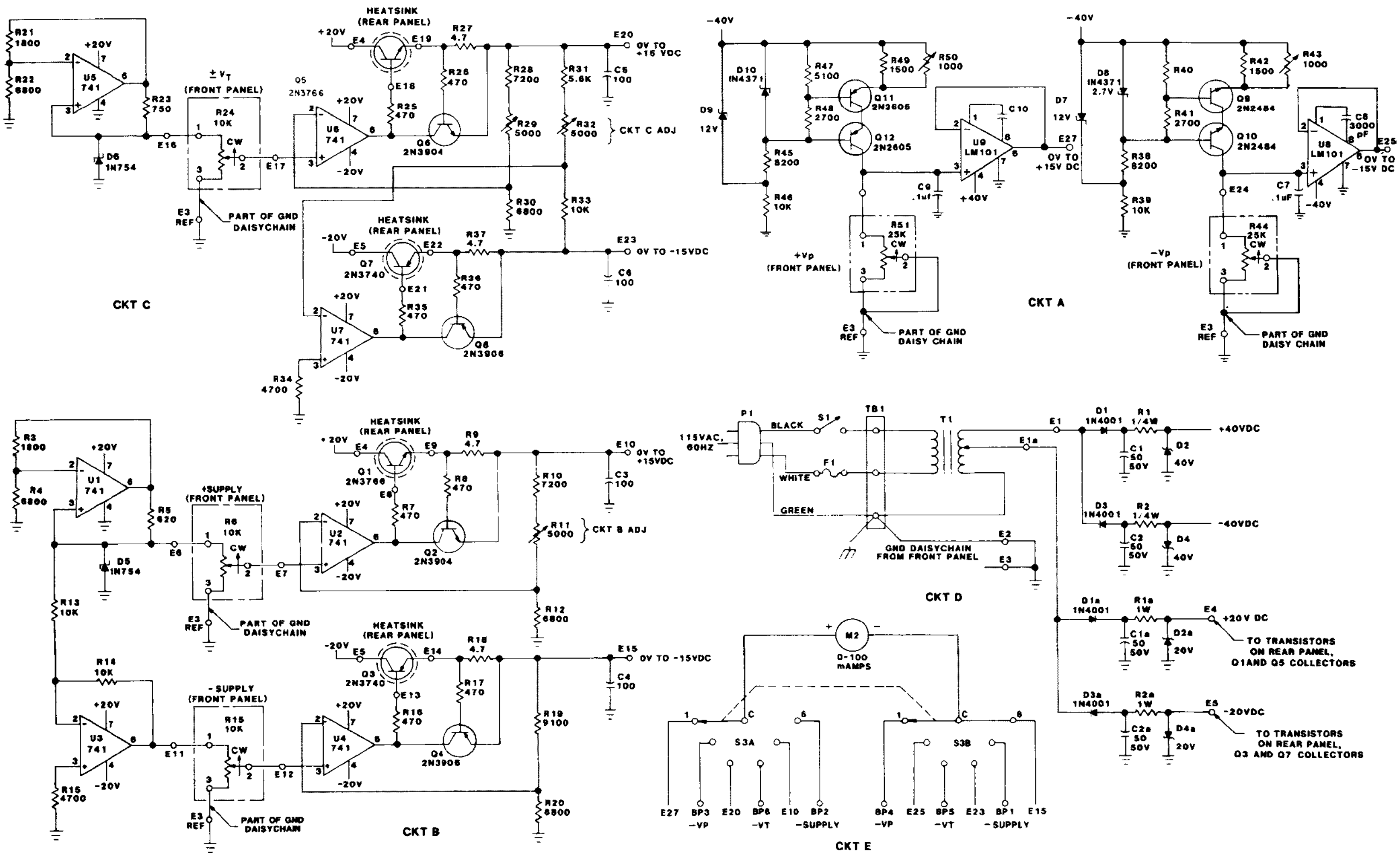


fig. 1. Six-output power supply schematic. In this schematic all capacitors are microfarads unless otherwise noted.

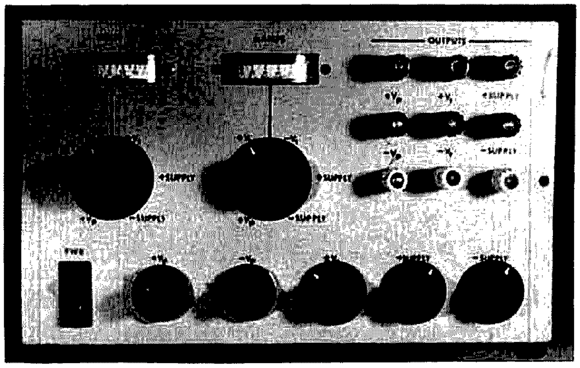


fig. 2. Front panel controls and indicators.

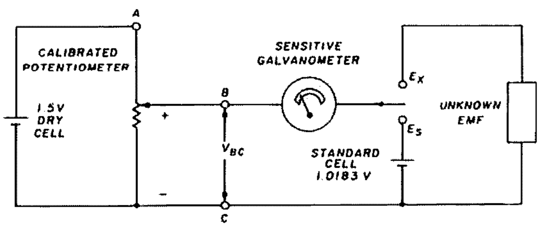


fig. 3. A typical application of a precision Weston electrochemical voltage standard cell.

limiting, and with the output series-pass transistors Q1 and Q2, providing the needed current drive, all five previously listed requirements of a voltage regulator are satisfied.

negative independently adjustable supply

The complementary negative portion of circuit B (CKTB) is composed of op amps U3 and U4 and associated circuitry. The reference voltage from the 6.8 volt zener diode D5 provides the input to an inverting amplifier (U3) with a gain of 1. The ratio of the input resistor (R13) divided by the feedback resistor (R14) establishes the gain, and since they are equal, the gain is 1. Therefore, amplifier U3 has a −6.8 volt output, which like its positive counterpart, is applied to front panel control pot R15 (− supply); refer to fig. 2. This negative input voltage is fed to the noninverting amplifier, U4, and amplified for a gain of from zero to approximately 2.5 as a result of the feedback resistor ratio from R19 to R20. The transistors Q3 and Q4 provide current buffering only — no voltage gain.

dual tracking power supply

This circuit, CKTC, is a composite and slight variation of previously developed and discussed circuits. Amplifier U5 has a +6.8 volt voltage reference in the form of zener diode D6 applied to the noninverting (+) input pin 3. This voltage is applied across front panel control pot ($\pm V_T$) R24. This 0 to +6.8 volts is then applied to noninverting amplifier U6, as described above. Amplifier U7 and transistors Q7 and Q8 form a negative regulator that is slaved to and therefore "tracks" the positive portion of this circuit, but naturally has an output equal in magnitude but opposite in polarity.

programmable power supplies

These two positive and negative power supplies are precision voltage sources and do not deliver power. They can deliver only approximately 4 mA of current, the capacity of an op amp itself. They are used primarily as precision stable voltage sources. Those readers old enough to remember the Weston cell that characterized most instrument labs some 20 years ago or more will recall that these were electrochemical voltage standards intended for use with very high resistance pots. They were typically employed with

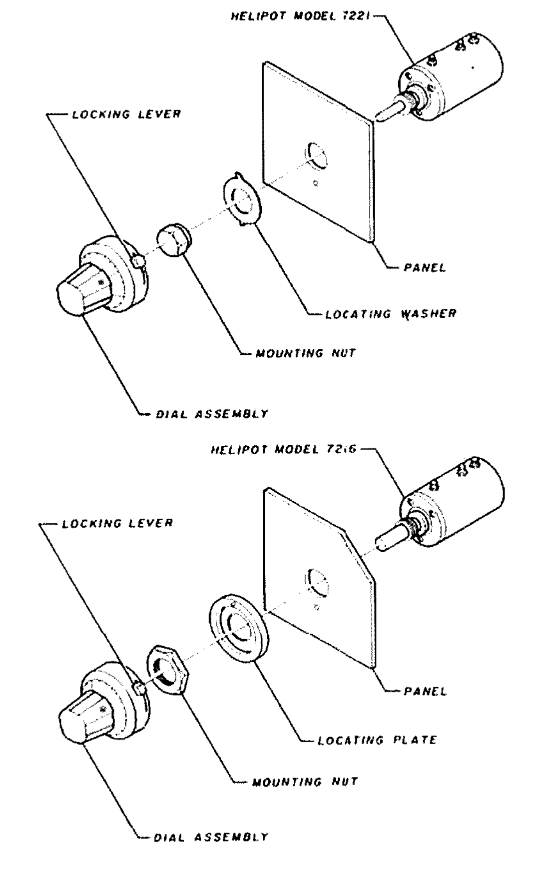


fig. 4. Mounting and functions of a multi-turn pot and matching turns counting dial.

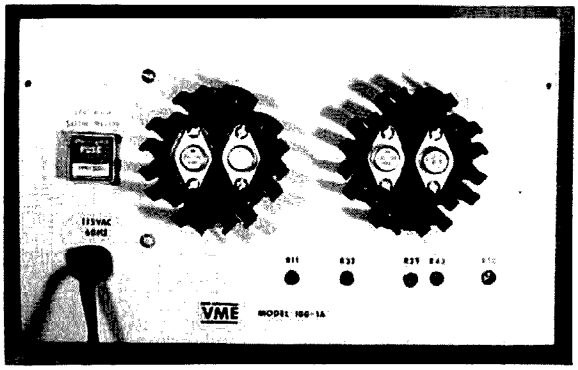


fig. 5. Rear view of case shows easy access to PC board trimmers.

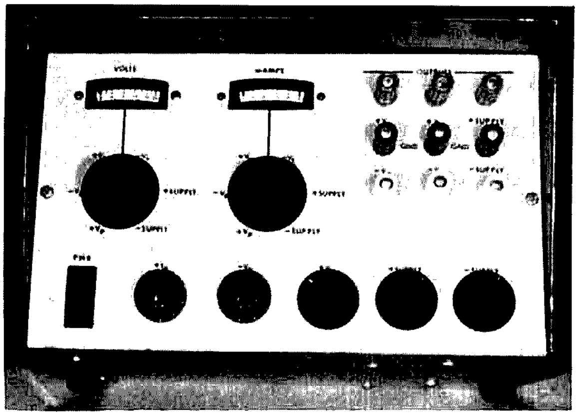


fig. 6. Six-output power supply.

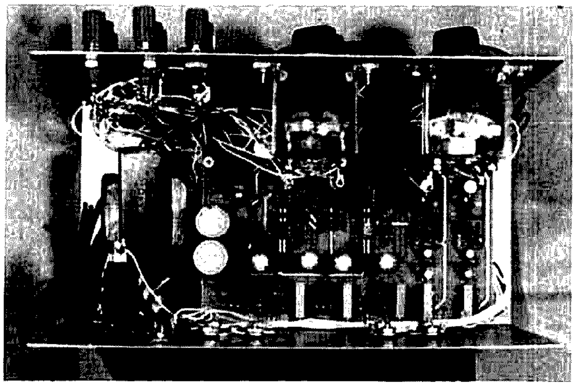


fig. 7. Interior view of power supply.

galvanometers in "balancing schemes" to indirectly determine to several significant figures the accuracy of an unknown voltage source (see **fig. 3**). Today we have 6 1/2-digit DVMs that will do this with equal or better accuracy and with much more convenience than that cumbersome hookup configuration. The Weston standard cells nonetheless developed an open-circuit voltage of 1.0183 volts as a result of the precise predictable nature of the electrochemical reactions within. But they could never be used to operate even a small lamp or similar low current demand device because irreparable damage would result, and their accuracy and repeatability could never again be trusted. However, these two precision low current circuits, using LM101 op amps, are nearly short-circuit proof.

The positive and negative programmable power supplies are simple (see **fig. 1**, CKTA). Both use complementary pair transistors to form an accurate current source that can provide 1 mA of current. This current, run through a 25-kilohm pot, generates a +25 volt fullscale output (positive power supply).

The pot, R51, is a 10-turn pot with a 10-revolution turns counting dial, (see **fig. 4**). For each revolution, up to and including the tenth turn, the pot will have 2.5 kilohms of resistance. Therefore, if you were to turn the pot to 4, you would have 4 turns at 2.5 kilohms each for 10 kilohms of resistance with 1 mA of current flowing through it for 10 volts DC of positive very precise voltage generated. You will note that this very precise positive voltage source goes to the input of U9 which is an op amp used as a voltage follower. More specifically, it is just an amplifier with a gain of +1 with a bit of current buffering. The noninverting input denoted by a (+) sign on U9 has an input impedance of at least 10 megohms, so very little, if any, voltage division occurs at this point.

The negative programmable power supply is the mirror image of its positive complementary counterpart. The +12 volt zener establishes bias on the bases of a complementary pair of transistors, Q9 and Q10, which produce a constant 1 mA current. This current is adjusted by a multi-turn PC mountable pot placed at the rear of the PC board (see **fig. 5**). Adjust R43 until four turns of the pot (10K) produce exactly 10 volts. A DVM can be used to check this. Make sure you do not load down this output. Op amp U8, like all nine op amps used with this project, is an LM101. If you have the TO-100 version, use it; its pin configurations is identical (pin 8 is the pin underneath the tab).

construction

The Bud box selected and specified in the parts list is ideal for this application in terms of cost and shape. The top of the enclosure slides down over the metal frame in a clamshell fashion. **Fig. 6** illustrates the finished product with the top in place. **Fig. 7** shows the project with the lid removed.

The actual detailed dimension of the box and the holes are provided in **fig. 8**. It should be noted that the two holes holding the shafts of the multi-turn pots can be either 1/8 or 1/4 inch, depending on the pot shaft diameter and corresponding turns counting dial

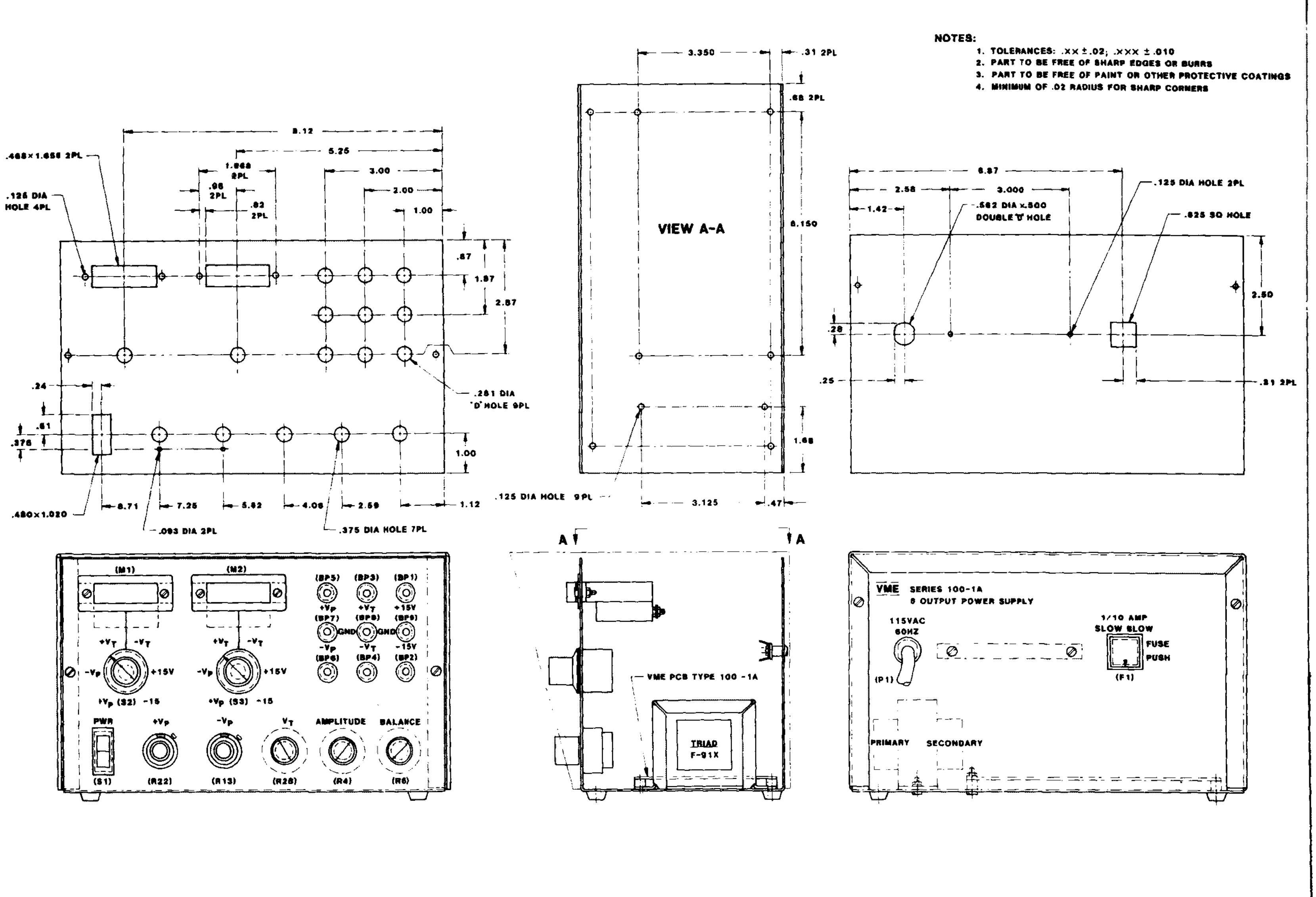


fig. 8. Drawing illustrates case drilling pattern.

parts list

resistors

item	description
R1,R2,R9,R18,R27,R37	4.7 ohms
R1a,R2a	4.7 ohms 1 watt
R3,R21	1.8 K
R4,R12,R20,R22,R30	6.8 K
R5	620 ohms
R6,R15a,R24	10 K front panel pot-single turn
R7,R8,R16,R17,R25 R26,R35,R36	470 ohms
R10,R28	7.2 K
R11,R29,R32	5 kilohm 15 turn 3/4" rectangular PC board mountable trimmer pot
R13,R14,R33,R39,R46	10 K
R15,R34	4.7 K
R19	9.1 K
R23	750 ohms
R31	5.6 K
R38,R45	8.2 K
R40,R47	5.1 K
R41,R48	2.7 K
R42,R49	1.5 K
R43,R50	1 kilohm 15 turn 3/4" rectangular PC board mountable trimmer pot
R44,R51	25 K ten-turn front panel pot

Note: all resistors are 1/4 watt, 5 percent tolerance unless otherwise noted.

capacitors

item	description
C1,C1a,C2,C2a	47 μF, 50 V
C3,C4,C5,C6	100 μF, 50 V
C7,C9	0.1 μF, 50 V
C8,C10	3,000 pF, 50 V ceramic dielectric

Note: all capacitors 47 μF and above require radial leads.

diodes

D1,D1a,D3,D3a	1N4001 or equivalent
D2,D4	40-volt zener or two in series to equal 40 V. A 39-volt zener (1N5366B) is available from Active Electronics; another, (1N4754A) is available from Circuit Specialists.
D2a,D4a	1N4747A
D5,D6	1N754
D7,D9	1N4742A
D8,D10	1N4371

transistors

Q1,Q5	2N3766 mounted on rear panel heatsink
Q2,Q6	2N3904
Q3,Q7	2N3740 mounted on rear panel heatsink
Q4,Q8	2N3906
Q9,Q10	2N2484
Q11,Q12	2N2605

op amps

U1 through U9	LM101 or equivalent

meters

Modutec edgewise 2-inch meters: part No. 13505 for a 1 mA (shunt for 100 mA); part No. 13537 for a 0-30 VDC meter. Both available from Sintec for \$10.20 and \$8.80 respectively. A 1 mA edgemeter (No. MET-1) is available for \$4.50 from All Electronics Corp. Another is offered by Circuit Specialists (No. 20-905) for \$6.07.

switches

item	description
2	6-position rotary switches
1	SPDT (single-pole double-throw) 115 VAC 1 amp

hardware

9	binding posts
1	fuseholder (a round one, it desired), part No. F150145-ND (\$2.31) or F004-ND (\$1.95), available from Digi-Key Another fuse holder (part No. FHPM-3), is panel-mounted, and fits 3AG-type fuses; it's available from All Electronics Corp.
2	matching 10-turns turns-counting dial (No. 411), \$11.50, available from Circuit Specialists
1	10k 10-turn pot (No. 73JA), \$7.50, available from Circuit Specialists
cabinet	Bud part No. SE3030

PC board, drilled and etched. Available from the authors for \$14.00 ppd. Contact Speciality Electronics Services, Inc., P.O. Box 3320, San Antonio, Texas 78211.

6' flat line cord, 18 gauge, 3-prong (No. FDAC-3) is available from All Electronics Corp.

Note: a 10k pot is usable on the two programmable voltage sources if a 0 to 10 VDC full-scale output is sufficient.

parts suppliers:

Active Electronics
P.O. Box 8000
Westborough, Massachusetts 01581

All Electronics Corporation
905 South Vermont Avenue
Los Angeles, California 90006

Circuit Specialists
P.O. Box 3047
Scottsdale, Arizona 85257

Digi-Key Corporation
Highway 32 South
Thief River Falls, Minnesota 56701

Sintec Electronics
28 8th Street
Frenchtown, New Jersey 08825

selected. Also, the cutout in the back, made with a Greenlee 5/8-inch square punch accommodates a rather fancy fuseholder. You may have greater luck finding a round standard MIL-M-14 type CFG fuseholder with a 0.440-inch diameter round hole. The 1:1 scale foil pattern are shown in **fig. 9.** Note that the side with far fewer runs (lands) is the *component* side. Kits are available (see **fig. 10**) for lifting a pattern directly from a magazine page such as this one. The drill diameter schedule shown in **fig. 11** is a must for this method; even if you purchase the pre-drilled board, you will nonetheless still need this drawing as a component placement guide.

When mounting the parts on the PC board, be sure to observe the polarity of all the diodes and polarized electrolytic capacitors. Also, be sure to note that the op amps have pin 1 indicated by a dimple adjacent to the notch. The PC board has a small dot above pin 1 on the foil side of the board. Failure to observe these guidelines will result in destruction of the op amps. The smaller TO-100 type of transistors that do not go on the rear panel in the heatsinks can be properly oriented onto the PC board by a "tick" mark denoting where the alignment of the case's tab should be. Diode symbols are on the board; there are also (+) signs for proper placement of electrolytics.

The meter circuit is optional because only one supply at a time can be used when the current meter is in use.

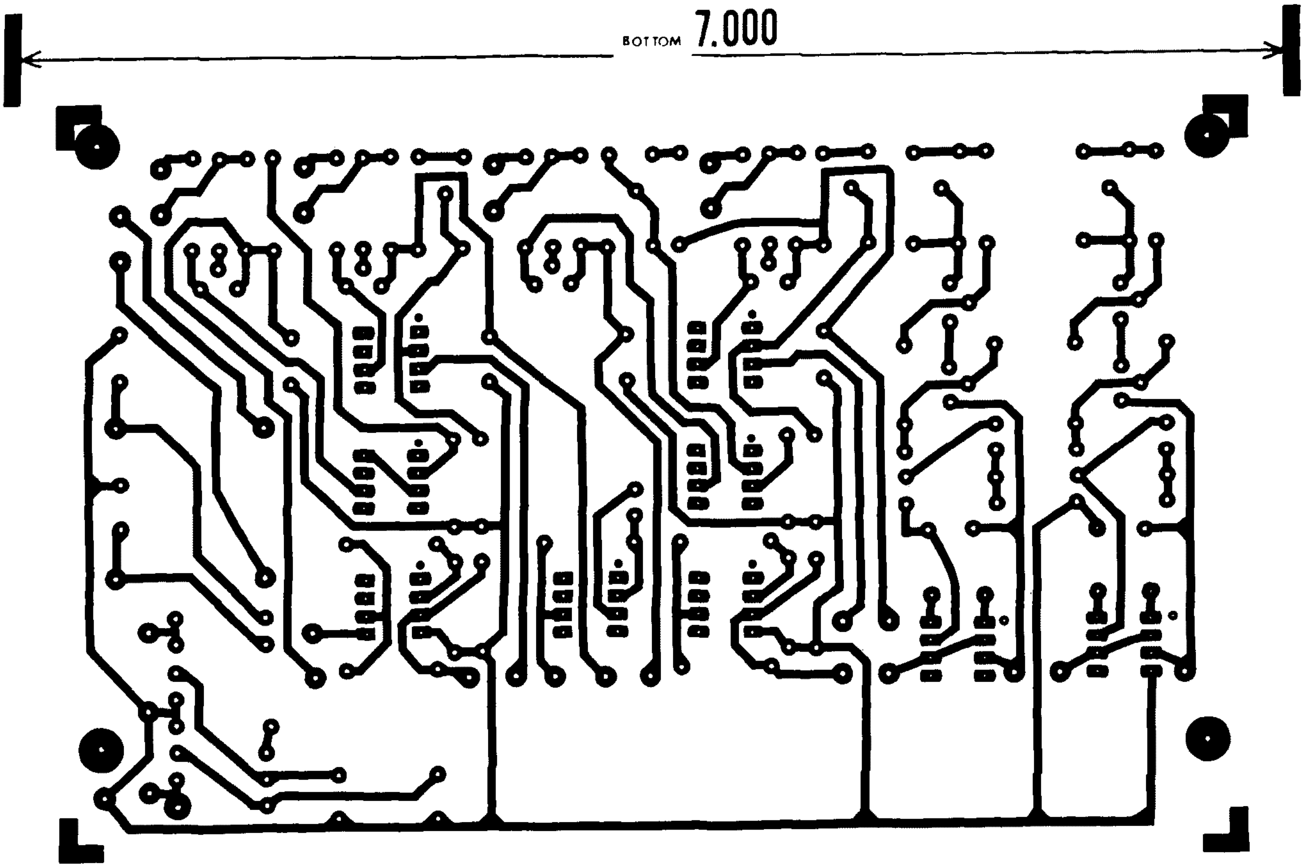
turns-counting dials

Refer to **fig. 4** when placing the turns counting dial and multi-turn pot on the front panel. Select a dial that matches the pot's shaft diameter; the two most common shaft diameters are1/4 and 1/8 inch. The turns counting dial that matches it will probably have its inner scale marked in gradients of hundredths of a turn. The outer scale counts the number of turns (usually up to 15) completed.

To mount the pots and dials, refer to **fig. 4** again and follow the procedure described below, prepared originally for the Helipot™ Duodial™ series of pots and multiple turns counting dials:*

- Locate positions for holes "A" and "B" on panel. "B" for lug on locating washer is 9/32 inch below center of hole "A".
- Drill 9/32 inch hole "A" in panel for pot bushing.
- Drill 5/64 inch hole "B" in panel for lug on locating washer.
- Turn potentiometer shaft against its counterclockwise stop. Insert shaft into center hole "A" in panel.

*Courtesy Beckman Instruments, Inc., 2500 Harbor Blvd., Fullerton, California 92634.



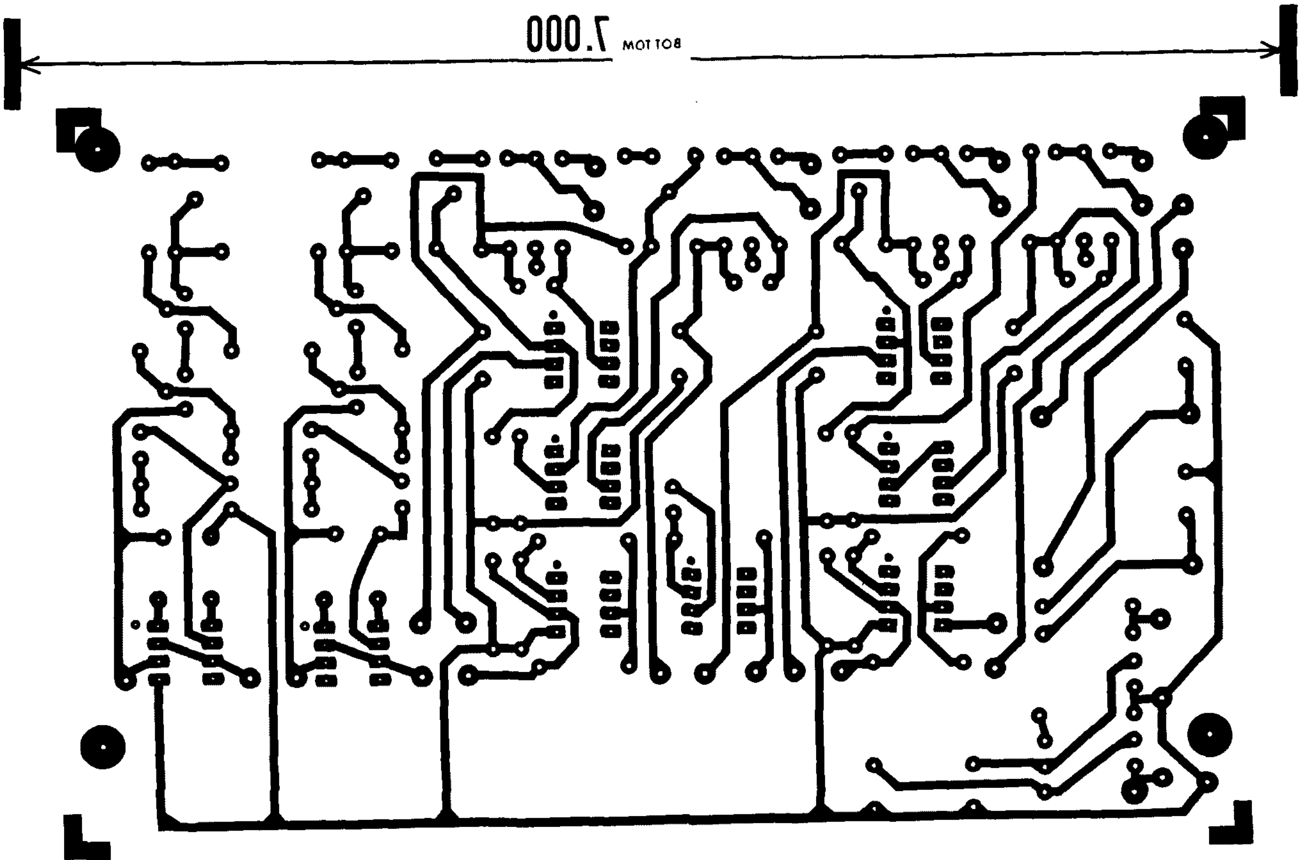
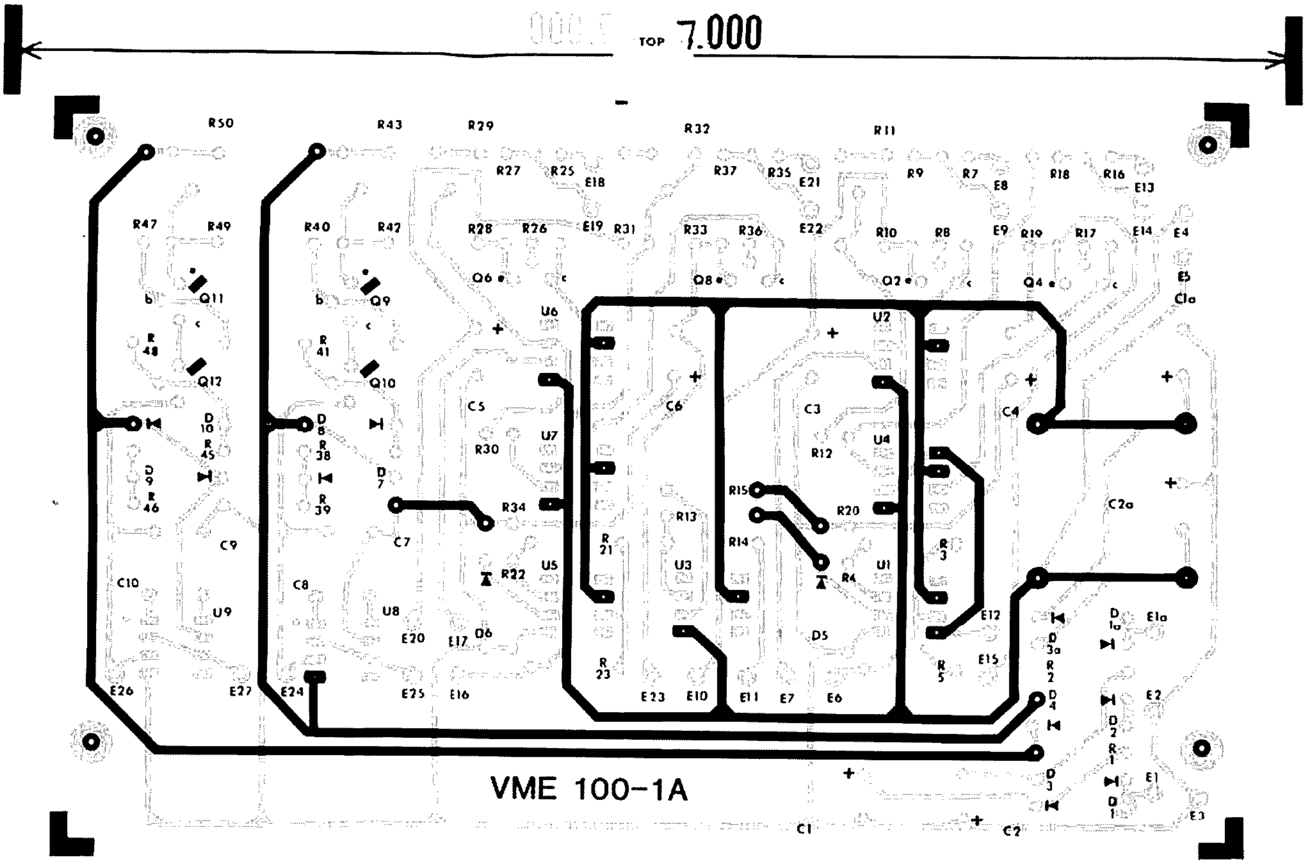


fig. 9. Double-sided PC board artwork.



• Slip locating washer over shaft and seat lug on locating washer in hole "B".
• With a wrench, firmly tighten mounting nut onto potentiometer bushing. Note that the nut supplied is reversible. For thick panels, use as shown in drawing. For thin panels, reverse the nut. If the Duodial is used with a device other than a Helipot potentiometer, an appropriate mounting nut may have to be obtained. This must fit within the 3/8-inch diameter 13/64-inch deep recess in model 2601 DUODIAL.
• With locking lever in OFF (UP) position, slip dial assembly over potentiometer shaft. Be sure lug at top of locating washer seats in slot in back of dial, and that the whole dial assembly sets lightly against the panel.
• Turn dial knob counter-clockwise until the zero of the outer scale is in the center of the window. Now turn slowly until the scale reads between 10 and 20 at the index line. Tighten the set screw until a very slight drag on the shaft is felt. Turn knob very slowly until both zeros line up with the index line. Tighten the set screw firmly.

parts — and where to get them

Components. This design purposely uses components that are easy to obtain. Several suggested sources are listed at the end of this article.

fig. 10. Typical kits available today to nonphotographically "lift" a PC board from magazine pages.

Zener diodes. Four are used in the input power conditioning circuit off the transformer. The two 20-volt zeners are 1N4747As, available from Digi-Key. The 40-volt zeners must be obtained elsewhere. However, two 20-volt zeners may be connected in series to achieve the same result. The other six low-power zener diodes consist of: two 2.7-volt zeners, two 6.8-volt zeners, and two 12-volt zeners. If the 2.7-volt zeners are hard to find, 1N5226B 3.3-volt zeners may be used in their place. If this substitution produces more than a 1 mA output from the constant current sources, then trim pots R43 and R50 can be adjusted accordingly for a 1 mA output. The two 12-volt zeners can

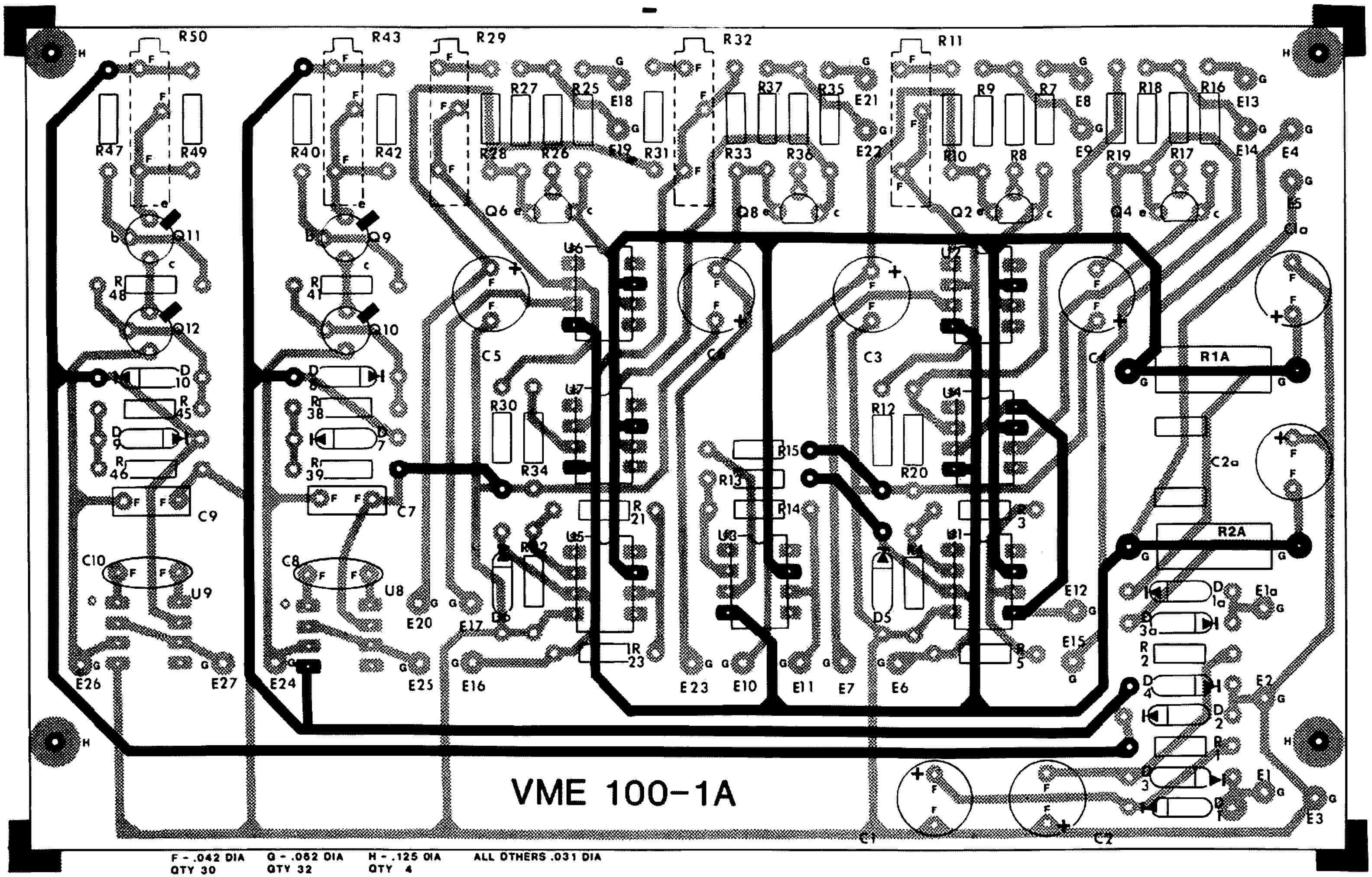


fig. 11. A combination drill pattern/component placement drawing.

be 1N963, 1N5242B, or 1N4742As. The 6.8-volt zeners establishing reference input voltages, namely D5 and D6, can be either 1N957B, 1N5235B, or 1N4736As.

Rectifier diodes. The rectifier diodes are 1N4001s. You may also use any of the 1N4000 series up to and including 1N4007, 1000 PIV diodes.

Capacitors. All capacitors that are polarized aluminum electrolytics are radial rather than axial lead devices, which are designed to lie flat on the PC board.

Meters. Edgewise meters have been selected to minimize the amount of front panel space required and to avoid a cluttered look.

Trim pots. All five cermet trimming pots are 3/4-inch rectangular ones with three staggered pins coming out their underside. All five are available from Digi-Key for $1.20 each. The three 5K pots are part No. 01B53 and the two 1K pots are No. 01B13.

Op amps. It is the nine LM101 or 107-type bipolar operational amplifiers that give this project its appeal in the form of some exceptional performance parameters.

National Semiconductor, the leader in op amp and linear IC manufacturing, has adopted a parts numbering convention now almost universally accepted throughout the industry, with variations occurring only in prefixes to denote the different styles of case available. "L" in the prefix "LM" stands for "linear," the "M" stands for "monolithic," (as opposed to "H", which stands for "hybrid"). The first digit — a 1, 2, or 3 — signifies whether the part is military, industrial, or commercial grade. A "3", or commercial grade product, is sufficient for this project and is less expensive than a grade 1 or 2 part.

Transformer. The Triad F-91X specified will adequately power this project and its ±100 mA outputs. However, if 2 amp outputs are required, you must increase the transformer's capacity along with the output power transistors' drive capability into 12 amp Darlington or single-cased direct-coupled transistor pairs. The two NPN transistors, which are now 2N3766's must then be 2N6057's or equivalents. The two PNP transistors, which are now 2N3740's, must then be 2N6050's or PNP Darlington pairs. Housed in TO-3 cases, both require heatsinking. Detailed thermal design information is available in references 1 and 2.

references

1. Vaughn D. Martin, "Cooling Semiconductors: Part One," *ham radio*, July, 1984, page 33.
2. Vaughn D. Martin, "Cooling Semiconductors: Part Two,": *ham radio*, August, 1984, page 52.

ham radio

Leyden's magic jar: the derivation of the Hertzian and Marconi antennas

From "electrical fire" to today's "rubber duckie"

In 1745, Ewald Jurgen von Kleist literally held in his hand the ancestor of most radio antennas in use today. Von Kleist was not a radio engineer; radio wasn't even discovered until 43 years later. Dean of the Cathedral of Kamin in Pomerania, Kleist enjoyed investigating the strange phenomenon of "electrical fire" in his spare time. Realize, now, that this was long before there was an electric light bulb, electro-magnet, or even a battery with which to power them. Von Kleist was not seeking to discover anything about antennas — he'd never even heard of them. He was just enjoying himself, as many experimenters do, by playing with and studying electricity. In doing so, he unwittingly discovered the world's first electrical "condenser" or capacitor. And believe it or not, this discovery led to antennas as we know them today.

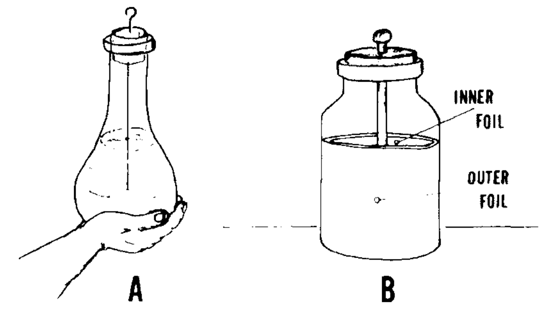


fig. 1. (A) An example of a capacitor such as von Kleist discovered; (B) A Leyden jar (occasionally called a "Kleistian jar).

As you know, a capacitor consists of two conductors called "plates" which are separated by an insulator called a "dielectric." Such a device can accept and store an electrical charge. First Kleist filled a glass flask with water. Then, holding the flask in one hand, he inserted a wire into the water in the flask and charged the wire with sparks of static electricity. As shown in **fig. 1A**, the water in the flask formed the inside plate of the world's first capacitor. The glass of the flask formed the dielectric, and von Kleist's hand served as the other plate of the capacitor! Touching the wire, von Kleist literally got the "shock of his life." He told others who played with electricity about his "shocking" discovery; one of them, a man named Graylath, succeeded in replicating the effect. According to Heilbron,[1] Graylath reported that his device for storing "electrical fire" could knock children of 8 or 9 "off their feet."[1] Such was the remarkable beginning of what we commonly refer to as the "Leyden jar." The term "Leyden" derives from the town of Leyden in the Netherlands, where in 1746, Pieter van Musschenbroek and his student Cuneaus further developed the early capacitor while trying to replicate and improve upon von Kleist's work. Obviously, one could not continue to use one's own hand as one plate of this potentially dangerous device. In the Leyden jar, the experimenter's hand on the outside of the flask was replaced with a coating of metal foil. The water on the inside of the flask — the inner plate of the capacitor — was

By W. Clem Small, KR6A, 26530 Parkside Drive, Hayward, California 94542

replaced by a second piece of foil. These two foil plates were separated by the glass wall of the Leyden jar, just as the glass of the flask had separated the water and von Kleist's hand in the original device. **Fig. 1B** shows the Leyden jar invented by Musschenbroek.

As the early physicists continued to pursue the elusive electrical phenomena of the day, more discoveries were uncovered. Ruhmkorff developed the induction coil or transformer. Induction coils equipped with the battery, invented by Volta, were capable of producing very high voltages continuously and giving impressive sparks across gaps in conductors. It was also found that the induction coil would give a much more powerful spark at a spark-gap if a Leyden jar or two were connected across the gap. Another finding quite important to the discovery of radio was that the discharge of a Leyden jar across a spark-gap was actually an oscillating current rather than just one pulse of current as it appears to the eye. Then along came Heinrich Hertz, bringing these discoveries together to confirm Maxwell's predictions of the possibility of radiating electromagnetic waves through space. As part of the apparatus used in demonstrating this remarkable phenomenon, he employed an induction coil and spark-gap to create a transmitter similar to that of **fig. 2**.

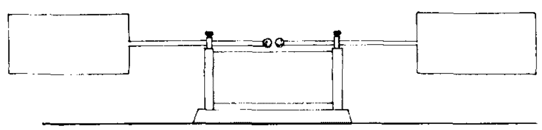


fig. 2. An example of a spark-gap transmitter with dipole antenna and capacitance plates.

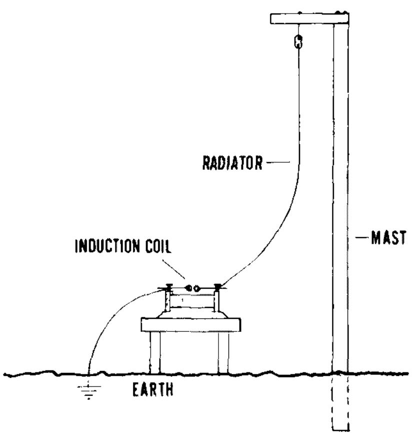


fig. 3. A Marconi antenna connected to a spark-gap transmitter. The left side of the spark-gap is connected to the earth.

Aitken has noted that the metal plates on the ends of Hertz's dipole antenna (**fig. 2**) were essentially Hertz's way of "unrolling" the foils of the Leyden jar and thereby increasing the capacitance across the spark-gap of his transmitter.[2] Hertz, of course, knew that a Leyden jar discharging through a spark-gap produced an oscillating electric current. What he had to do was to somehow "open up" that circuit so that it could produce the suspected radiation of electromagnetic waves that Maxwell had predicted. In discussing the evolution from Leyden jar-plus-wire-loop oscillator to wire-plus-metal-plate antenna, Aitken says, "The two foil surfaces of the Leyden jar, opened outward and transformed, became Hertz's radiating dipole antenna."[2] Hertz at times dispensed with the foils, and used as his antenna only the rods to which the foils were attached. (In some instances he selected dipole lengths that put his electromagnetic radiations near the 2-meter Amateur band of today!) Thus did the Leyden jar evolve into the dipole or doublet antenna as we now know it.

Subsequently, Guglielmo Marconi, while vacationing in the Italian Alps, read an announcement of the death of Hertz. The same article recounted Hertz's astounding achievements in demonstrating the possibility of transmitting electromagnetic waves through space. Marconi was struck with the potential such a discovery might hold for the communication of information across more than just the few feet of laboratory space which Hertz had done. He cut short his vacation and returned home to begin work on what was to become the wireless telegraph, or what we now call radio. Marconi soon began experimenting with antenna placement and eventually came to connect one side of the spark-gap in his transmitter to a vertical antenna (see **fig. 3**) and the other side of the gap to the earth or ground.

Hertz had found that he had to make the length of his dipole equal to one-half the wavelength of the frequency which he desired to radiate. Marconi, on the other hand, found that with the earth substituting as half the dipole, he could use one-fourth of the wavelength as the length of his vertical radiator. With this configuration was born the grounded vertical antenna, ancestor to the common quarter-wave vertical ground-plane antenna as well as AM broadcast towers and many other vertical antennas. Although antennas didn't look much like a capacitor or a Leyden jar by Marconi's time, the idea of the antenna as a capacitor did not die with Hertz's opened-out Leyden jar. Early technical books on radio routinely discussed antennas as being large capacitors and explained their functioning in those terms. For instance, in a 1922 revision of the U.S. Army Signal Corps radio handbook,[3] we find the following: "There are two general types of anten-

nas, those which act primarily as electrical condensers and those which act primarily as electrical inductances. The first type is usually referred to simply as an 'antenna.' The second type is usually referred to as a 'coil antenna,' 'coil aerial,' 'loop,' or when used for a particular purpose, as a 'direction finder.' " Of course today we realize that an antenna has capacitance (as well as inductance and resistance), but are more likely to conceptualize an antenna as a resonant circuit than as a capacitor.

terminology

Notice that in fig. 2, Hertz's antenna is oriented with its radiating conductor in the horizontal plane. Thus the electron flow on the radiator, and the electrical wave radiated into space, would naturally be horizontal. For this reason, Hertz's antenna is described as a horizontally polarized antenna. Terms still in use for common horizontally polarized antennas, in particular the $1/2\lambda$ dipole or doublet, are "Hertz" or "Hertzian" antennas. On the other hand, in Marconi's $1/4\lambda$ antenna, note the vertical orientation of the antenna's wire radiator. Here the current flow on the radiator, and the electrical wave radiated into space, are vertically oriented. It's not hard to see why, even today, vertically polarized antennas worked against a groundplane are sometimes referred to as Marconi antennas. For example, in the 1982 edition of the RSGB's *HF Antennas*, Moxon refers to ". . . the Marconi antenna . . . which can be regarded as one-half of a dipole, the other half being the image in the ground."[4] Shrader, in *Electronic Communication*,[5] wrote, "Any antenna complete in itself and capable of self-oscillation, such as a half or full wavelength, is known as a *Hertz* antenna. When an antenna utilizes the ground (earth) as part of its resonant circuit, it is a *Marconi* antenna." To me, it seems fitting that the two most common antenna configurations we have today still bear the names of their illustrious discoverers.

The next time you whip the ol' rubber duckie past your ear as you put your HT on the air, give a little thought to the debt we owe those old-timers for the pleasures we enjoy today. If the thought makes the hair on the back of your neck stand up a little, it might just be that the spirit of old von Kleist is back there with a charged flask, still playing with his "electrical fire."

references

1. J.L. Heilbron, *Electricity in the 17th and 18th Centuries*, University of California Press, 1979, Chapter XIII.
2. H.G.J. Aitken, *Syntony and Spark*, John Wiley, New York, 1976, pages 53 and 56.
3. *The Principles Underlying Radio Communication*, 2nd edition, U.S. Army Signal Corps, 1922, page 304.
4. L.A. Moxon, *HF Antennas for All Locations*, Radio Society of Great Britain, 1982, page 103.
5. R.L. Shrader, *Electronic Communication*, 2nd edition, McGraw-Hill, New York, 1967, page 439.

ham radio

low-cost UHF antenna tower

Chimney-mounted perforated tubing provides accessible 57-foot skyhook

This article describes a design for a UHF antenna tower that can be constructed very simply and at a price most hams should be able to afford — less than $200.00. This type of tower is ideal for Amateur Fast Scan Television applications, among others.

Before the design was begun, several ground rules were established:

- The tower should be attached to the chimney of the house. (The advantage is obvious; the chimney would serve as the base for the tower and also provide extra height.)

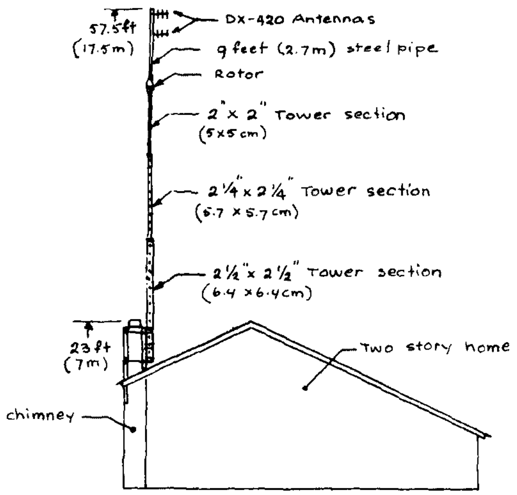


fig. 1. Tower and antenna assembly measures 57.5 feet (17.5 meters) from ground.

- For aesthetic reasons, no guy wires were to be included. This meant that the tower had to be strong enough to withstand 80-mile per hour (128 km per hour) winds with an antenna area of 2 ft^2 (0.19 m^2), which is sufficient for two or *more* UHF-type antennas.
- For ease of maintenance, the tower must be a tiltover.
- The tower sections had to be light enough to allow construction and assembly to be carried out without the need for a crane.

the chimney must take the load

The first order of business was the necessity of establishing the strength factor of the chimney for a given side load. This is very important; not all chimneys are strong enough to support a mast. At this site, the chimney is constructed from an inner column of tube segments surrounded with a brick-cemented outer shell. The space between the brick shell and the tube segments is filled with concrete reinforced with four 1/2-inch (1.27-cm) steel rods. This is common practice in Southern California. A check with the county masonry society revealed that this type of chimney should be capable of sustaining a side load of 1000 pounds (454 kg).

The next step requires calculating the strength specifications for the tower itself. Because no guy wires were planned to support the tower above the chimney, the combined strength of the tower, the rotator mounted on top, and the antennas above the rotator must be sufficient to sustain violent winds up to 80 miles per hour, or 128 km per hour. (A detailed analysis is included at the end of this article.)

The tower and antenna assembly consists of a set of chimney mounting brackets; three lengths of square tubing sections, each successively smaller in diameter and fitting inside the previous tube; a rotator; and a

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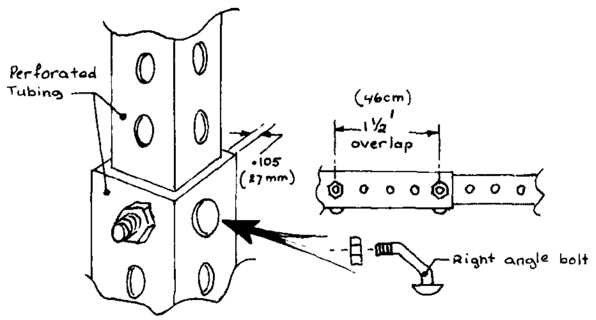

fig. 2. Right angle bolts connect tubing sections.

9-foot (2.74 meters) long, 16-gauge (2 mm) galvanized steel 1-5/8 inch (4.1-cm) I.D. tube with two DX-420 Cushcraft antennas mounted to it. The total height, measured to the tip of the antenna mounting tube is 57.5 feet (17.5 meters) above the ground. This includes the 23-foot (7-meter) high chimney. (See **fig. 1**).

construction details of tower and chimney mounts

The tower was assembled from three lengths of Telespar perforated tubing. This type of tubing was designed for sign posts, storage racks, and benches. Of the many unique properties of the tubing, its modularity makes it most useful for antenna mounting. Each segment of tubing was designed to fit into the next larger tube diameter. Right angle bolts provided by the supplier fasten the sections of tubes together. (See **fig. 2**). An overlap of 1-1/2 feet (46 cm) should be maintained to allow for sufficient strength. Since cost and simplicity were of primary concern, no elaborate motor driven or other type of mechanism to lower or raise the tower was included. Access to the antennas was achieved by making it a tiltover assembly with a hinge.

The brackets are of an all-welded construction. A detailed drawing (**fig. 3**) outlines the construction features of the front bracket which supports the tower.

The rear bracket is almost the same as the front bracket, with the exception that no tower brackets are included. The author chose to make the bracket longer, reaching further down the chimney for added strength and to provide a better stress distribution in the chimney. Each chimney width will be different; thus the inside bracket dimension will be unique for each installation. The 4.5-foot (137-cm) bracket height should be maintained in order to provide adequate base support. Remember, the tower, rotator, boom, and antennas weigh approximately 150 pounds (68 kg). The front bracket weighs approximately 80 to 100 pounds (36 to 46 kg) and the rear bracket approximately 70 to 90 pounds (32 to 41 kg).

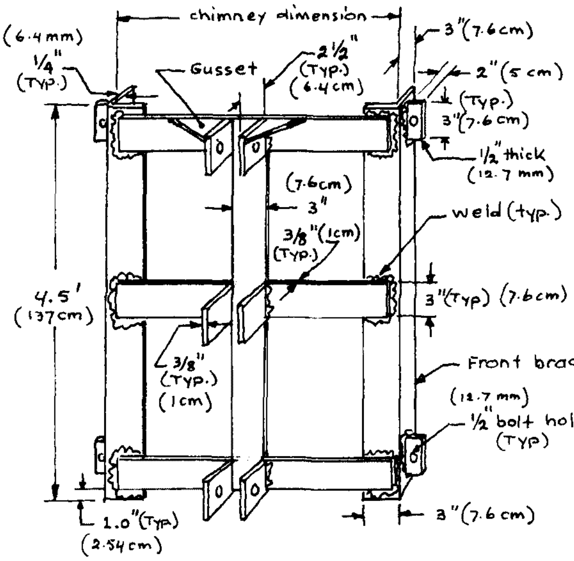

fig. 3. Construction details of front bracket assembly.

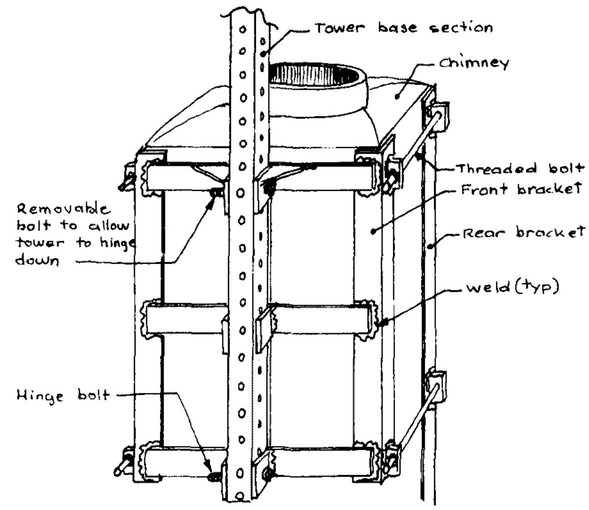

fig. 4. Hinge assembly, consisting of a front and rear bracket, is fastened with two 2-foot long threaded bolts.

After the brackets are completed, apply rust preventing paint and measure the width of the chimney accurately; the welded bracket assembly cannot be readjusted to fit the chimney. (A bolted construction may work loose or cause twisting in heavy winds.)

The sandwich mounting of the front and rear brackets, using 1/2-inch (12.7-mm) bolts, serves two purposes: first, it eliminates the need for drilling into masonry and mounting of lag bolts, which are unreliable in this application; second, it helps prevent the failure of masonry between the bricks. If the chimney is not straight you may have to insert shims between the brackets and the chimney for stability.

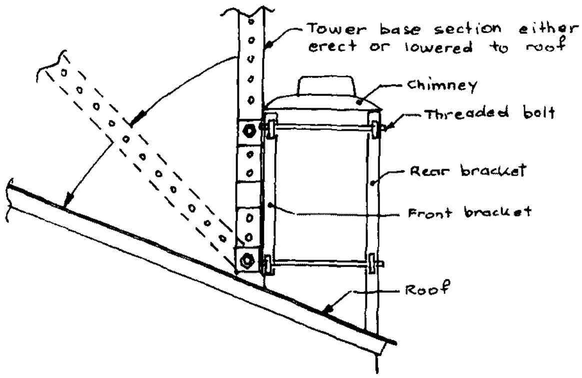

fig. 5. Remove upper mast bolt to lower tower to roof.

parts list

tower parts

quantity	part number	description
1	24F12*	perforated tubing 2-1/2 × 2-1/2 inches (6.4 × 6.4 cm), wall thickness 0.105 inch (2.7 mm), 12-feet (3.7 m) long, pre-galvanized steel
1	22F12*	perforated tubing 2-1/4 × 2-1/4 inches (5.7 × 5.7 cm), wall thickness 0.105 inch (2.7 mm), 10-feet (3 m) long, pre-galvanized steel
1	20F12*	perforated tubing 2 × 2 inches (5 × 5 cm), wall thickness 0.105 inch (2.7 mm), 10-feet (3m) long, pre-galvanized steel
2	TL070*	right angle bolt, electro-galvanized finish
2	TL050*	right angle bolt, electro-galvanized finish
1		tubing 1-5/8 inch (4.1 cm) ID, wall thickness 1/16 inch (1.6 mm), 9-feet (2.7 m) long

Telespar parts can be obtained from the Unistruct Corporation, 14600 South Marquardt Avenue, Santa Fe Springs, California 90670.

bracket parts

quantity	description
2	angle iron 3 × 3 × 1/4 inch (7.6 cm × 7.6 cm × 6.4 mm), 52 inches (1.37 m) long
2	angle iron 3 × 3 × 1/4 inch (7.6 cm × 7.6 cm × 6.4 mm), 6 to 8 feet (1.8 to 2.4 m) long
6	flat cold rolled steel plate 3 inches (7.6 cm) wide, 3/8 inch (1 cm) thick, length dependent on chimney width
6	flat cold rolled steel plate 3 × 2-1/2 × 3/8 inches (7.6 × 6.4 × 1 cm)
8	flat cold rolled steel plate 3 × 2 × 1/2 inch (7.6 cm × 5 cm × 1.27 cm)
2	galvanized steel bolt 5 inches (12.7 cm) long × 1/2 inch (1.27 cm) thick
8	galvanized steel nut for 1/2 inch (1.27 cm) bolt
4	galvanized threaded bolt 3 feet (91 cm) long — dependent on chimney dimensions
1	flat cold rolled steel plate 3 inches (7.6 cm) wide, 3/8 inch (1 cm) thick, 52 inches (1.3 m) long

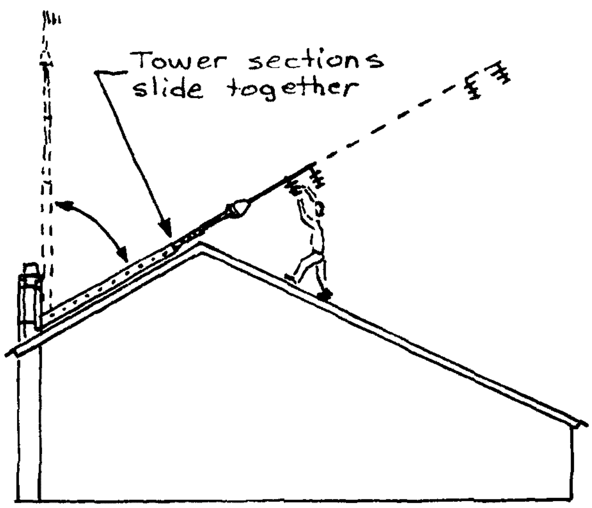

fig. 6. To reach antennas, simply remove bolts joining sections, sliding one section inside the other.

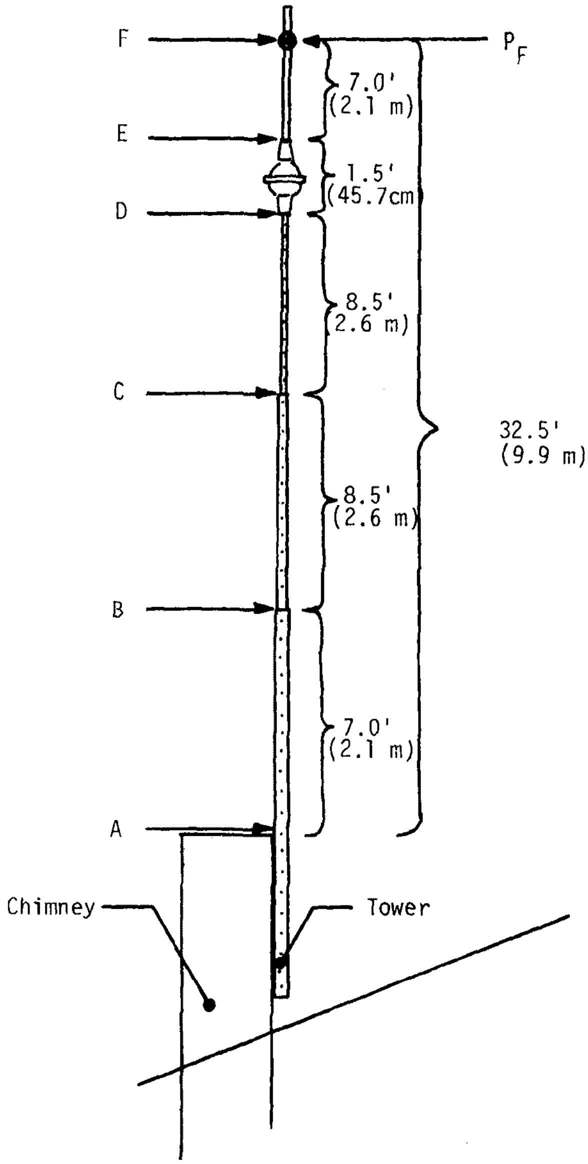

fig. A1. Tower/antenna configuration.

Designed to be part of a welded bracket straddled around the chimney, the hinge is fastened together with 1/2-inch (12.7-mm) two-foot long threaded bolts available at most hardware stores. (See **fig. 4.**) **Fig. 5** illustrates how the tower can be lowered to the roof by removing the upper mast bolt.

If you have a pitched roof, how do you reach the antennas after the mast is hinged down? The mast is 39.5-feet (12-meters) long, measured from the hinge point. It overhangs the rooftop by 27 feet (8.2 meters). *Therefore, the design includes the* sectional feature shown in **fig. 6.** Since three square tubes slide together, merely removing the two sets of bolts that hold the sections together allow shortening the mast

to the point where the antennas can be easily reached.

In terms of the rotator, antennas, and tubes supporting *these antennas, each individual must* consider the options available and choose the most appropriate configuration. The tower constructed by the author was tested involuntarily when *seven* major storms swept through the Los Angeles area with wind velocities up to 100 miles per hour (160 km) during the winter of 1982-1983. The tower, brackets, and antenna sustained no damage; the tower is still perfectly vertical, and no stress cracks or other minor damage has been observed on the brackets or tower.

acknowledgement

My sincere thanks to Bob Provost for providing the mechanical analysis for the tower design.

appendix

antenna tower stress analysis

The following is a stress analysis of an antenna tower attached to a chimney. This analysis may be used as a guideline for any tower configuration. The simplified approach assumes a cantilever beam with a projected antenna surface area of 2 feet^2 (0.19 meter^2). The object of the analysis is to determine the wind velocity a given tower/antenna configuration can endure without damage.

The configuration of the tower/antenna model is shown in **fig. A1.**

First, one needs to determine the moment of inertia (I_{SQ}) of sections "*A*" through "*F*," using equation:

$$I_{SQ} = 1/12\ (O.D.^4 - I.D.^4) \qquad \textbf{(A1)}$$

$$I_A = 1/12\ [(2.50)^4 - (2.29)^4] = 0.9635\ in.^4$$

Next, the *bending stress* (σ) as a function of wind force (P) needs to be determined.

$$\sigma = \frac{MC}{I} \qquad \textbf{(A2)}$$

where M = moment (arm)
C = distance to neutral axis of cross section
I = moment of inertia

For section "*A*" the bending stress is:

$$\sigma_A = \frac{(32.5\ ft)\ (P\ lb)\ [(1/2 \cdot 2.50\ in)]\ (12\ in/ft)}{0.9635\ in^4}$$

$$\sigma_A = 506.0\ (P)\ lb/in^2$$

The following table summarizes the results:

section	CG distance	tower cross section (inches)	wall thick (inches)	I	σ PSI
F	—	—	—	—	—
E	7.0	1-5/8 × 1-5/8	1/16	0.1592	428.7 P
D	8.5	*	—	—	—
C	17.0	2 × 2	0.105	0.4778	426.9 P
B	25.5	2-1/4 × 2-1/4	0.105	0.6925	497.1 P
A	32.5	2-1/2 × 2-1/2	0.105	0.9635	506.0 P

*Extra strong section used, calculation not necessary.

As can be seen from the table, Section *A* experiences the largest stress for a given wind force (Section *C*, the least). (In other words, Section *A* would be the first section to break.) The maximum wind force a section can withstand is based on the strength of the material.

The yield strength of the tower material is $\sigma\ yield = 33\ ksi$ (33,000 lb/in^2). Any bending stress less than 33 ksi applied to the cross section will *not* permanently deform the tower section.

The ultimate strength of the tower material is $\sigma\ ultimate = 52\ klb/in^2$. Any bending stresses from 33 to 52 klb/in^2 applied to the cross section may result in permanent deformation. Stresses above 52 klb/in^2 will cause the tower section to break.

Solving for the unknown wind force yields:

$$506.0\ P = 33\ klb/in^2$$

$$P = 65.22\ lbs$$

Wind force is related to wind velocity as follows:

$$P = \frac{A \varrho V^2}{g} \qquad \textbf{(A3)}$$

where A = antenna area
ϱ = density of air (0.076 lbm/ft^3 at 66°F)
V = wind velocity
g = gravitational acceleration

$$P = \frac{(2ft^2)\ (0.076\ lbm/ft^3)\ [(V\ miles/hr)\ \frac{1}{3600}\ \frac{hr}{sec}\ (5280\ ft/mile)]^2}{32.186\ [ft/sec^2\ (lbm/lbf)]}$$

$$P = 0.01016\ V^2\ lbs \text{ or } V = \sqrt{\frac{P}{0.01016}}$$

Therefore, the maximum wind velocity cross section "*A*" can withstand is:

$$V_{yield} = \sqrt{\frac{65.22}{0.01016}} = 80\ mph$$

and

$$V_{ultimate} = \sqrt{\frac{103.64}{0.01016}} = 101\ mph$$

Although the simplified calculations show a V_{yield} of 80 mph, winds up to and over 100 mph have not resulted in tower material yield at this location. It appears that using the equations outlined above should provide a conservative means of calculating the strength of a tower.

(The same analysis in metric units is available. Send an SASE . . . Editor.)

ham radio

John Marshall Haerle, WB5IIR

The Amateur Radio community was recently saddened to learn of the untimely death of John Haerle, WB5IIR, in an automobile accident August 1.

A popular author and lecturer active on 160 meters, John had been affiliated with Gates Radio as chief engineer for sales, and later with Collins Radio, where he served as director of advertising and public relations and also headed that organization's broadcast division.

Surviving are his wife, Rose; a son, Dan; and a granddaughter.

A scholarship fund has been established in John's name at North Texas State University; checks may be made payable to the North Texas State University Scholarship Fund and sent to the Dallas Amateur Radio Club, P.O. Box 173, Dallas, Texas 75220.

— ham radio

design your own trap antenna

Operate wherever you want with this efficient antenna

Even though the trap antenna seems to be simple, designing one — on paper or by cut-and-try — can be tedious. It's possible, however, to take the tedium out of the process by using a home computer or a programmable calculator. The antenna described in this article should perform acceptably on first use, and should be easy to adjust.

The basic concept underlying the trap vertical, dipole, or beam antenna is simple: the bottom (or half center) section is simply 1/4 wavelength long at the highest frequency. A parallel-resonant trap tuned to this operating frequency effectively isolates the rest of the antenna. On the next lower band, the bottom section will be less than 1/4 wavelength long (physically). This length is electrically extended by the loading coil effect of the below resonance trap (looks like an inductance), and the combination must be brought to 90 degrees effective (electrical) length by a second section of tubing. A trap resonant at this lower frequency can now be added if more bands are needed, or the antenna simply "stopped." Each additional band requires a trap, plus an extended metal section. At low frequencies, a loading coil can also be added to reduce section length.

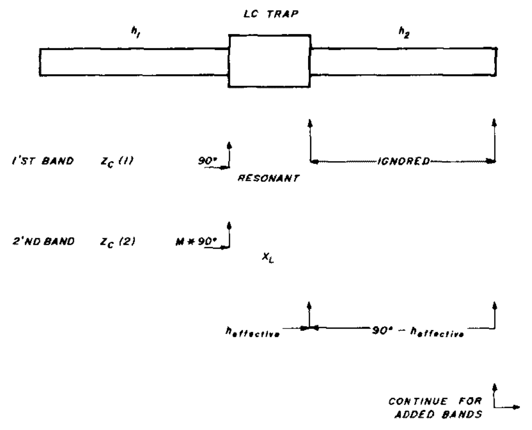

fig. 1. Organized approach to trap antenna design.

The challenge of trap antenna design lies in the fact that conversion from the physical length of sections to electrical length depends on the ratio of section diameter to wave length, so that the values must be recalculated for *each* band. Also, the trap-to-loading coil effect must be redone for each band, with conversion from reactance to effective length also considered. Unless an organized approach is used, it's easy to get lost.

For Amateur Radio purposes, the key to this organized approach is a series of articles by Boyer, W6UIH. The following summarizes the calculation steps involved and the equations used. (For explanations of the individual steps, refer to the Boyer series.)

INPUT the design parameters:

- Number of bands
- Operating frequencies, f_n, (highest first, in order)
- Section diameters, D_n
- Trap capacity or inductance, C_n or L_n

CALCULATE for specified frequencies:

- Trap inductance or capacitance, and reactance, X_n
- Section characteristic impedance, Z_0

CALCULATE by looping:

- Set first section length h to 90 degrees
- Loop for each band and for each trap
- Calculate the wave length = 11808/f inches
- Calculate the section impedance $Z_n = 60\ (L_n \cdot \lambda_n / D_n - 1)$
- Calculate the ratio of trap frequency to operating frequency $f_n/f_n' = m$
- Calculate effective section length $h' = mh$
- Calculate normalized trap reactance $X_L' = X_L\ (1/m - m)/Z_n$
- Add reactances $X' = X_L' + X$
- Convert to length $h = tan^{-1}\ X'$
- Determine section length when all traps and sections have been considered.

$S = 90° - h$

By R.P. Haviland, W4MB, 1035 Green Acres Circle, N., Daytona Beach, Florida 32019

```
  10 REM TRAP ANTENNA
  20 FAST
  30 PRINT "ENTER NUMBER OF BAND
S"
  40 INPUT J
  50 DIM F(J)
  60 DIM D(J)
  70 DIM L(J)
  80 DIM C(J)
  90 DIM T(J)
 100 DIM H(J)
 110 DIM Z(J)
 200 PRINT "ENTER TRAP KNOWN,L,C
,OR T IF COAX"
 210 INPUT K$
 220 PRINT "STARTING WITH HIGHES
T BAND,IN      ORDER,ENTER"
 230 FOR N=1 TO J
 240 PRINT AT 6,0;"FREQ ";N,
 250 INPUT F(N)
 260 PRINT F(N),"SECTION DIA ";N

 270 INPUT D(N)
 280 PRINT D(N),"TRAP ";N;" L-UH
,C-PF","OR 0 FOR COAX",
 290 INPUT TEMP
 300 IF K$="L" THEN LET L(N)=TEM
P
 310 IF K$="C" THEN LET C(N)=TEM
P
 320 IF K$="T" AND TEMP<>0 THEN
GOTO 280
 330 PRINT TEMP
 335 SCROLL
 336 SCROLL
 337 SCROLL
 338 SCROLL
 339 SCROLL
 340 SCROLL
 350 NEXT N
 360 IF K$<>"T" THEN GOTO 600
 400 REM COAX TRAP
 410 PRINT AT 10,0;"ENTER TRAP D
IA, ONLY 7/8 OR 1.5 (INCH FORM D
IA.USING RG-58 COAX)"
 420 INPUT TRP
 430 FOR N=1 TO J-1
 440 IF TRP=7/8 THEN LET T(N)=13
5.7*F(N)**-0.91
 450 IF TRP=1.5 THEN LET T(N)=68
.9*F(N)**-0.86
 460 IF NOT (TRP=7/8 OR TRP=1.5)
 THEN GOTO 410
 470 LET L(N)=(TRP+0.2)**2*T(N)*
*2/(18*(TRP+0.2)+8*T(N))
 480 NEXT N
 600 REM DO L OR C
 610 FOR N=1 TO J-1
 620 IF K$="L" OR K$="T" THEN LE
T TEMP=2*PI*F(N)*L(N)
 630 IF K$="C" THEN LET TEMP=10*
*6/(2*PI*F(N)*C(N))
 640 IF K$="L" OR K$="T" THEN LE
T C(N)=10**6/(2*PI*F(N)*TEMP)
 650 IF K$="C" THEN LET L(N)=TEM
P/(2*PI*F(N))
 660 NEXT N
 800 REM DO SECTION Z
 810 FOR N=1 TO J
 820 LET TEMP=11808/F(N)
 830 LET Z(N)=60*(LN (TEMP/D(N))
-1)
 840 NEXT N
1000 REM MAIN LOOPS
1005 LET DEG=0
1010 LET H(1)=PI/2
1020 FOR N=2 TO J
1030 FOR P=1 TO N-1
1040 LET M=F(N)/F(P)
1050 LET DEG=DEG+H(P)*M
1060 LET REACT=2*PI*F(P)*L(P)/((
1/M-M)*Z(N))
1070 LET REACT=REACT+TAN DEG
1060 LET DEG=ATN REACT
1090 NEXT P
1100 LET H(N)=PI/2-DEG
1110 LET DEG=0
1120 NEXT N
2000 REM OUTPUT
2010 CLS
2020 PRINT "FREQ DIA LUH CPF HDG
  HIN "
2030 IF K$="T" THEN PRINT AT 0,2
7;" TRNS"
2040 FOR N=1 TO J
2050 PRINT TAB 0;F(N);
2060 PRINT TAB 5;D(N);
2070 PRINT TAB 8;(INT INT (10*L(
N)))/10;
2080 PRINT TAB 13;INT (C(N)+0.5)
;
2090 PRINT TAB 17;(INT (10*H(N)*
180/PI))/10;
2100 LET FACT=0.673*Z(N)**0.053
2110 LET FACT=FACT*11808/F(N)
2120 LET FACT=FACT*H(N)/2/PI
2130 PRINT TAB 22;(INT (10*FACT)
)/10;
2140 IF K$="T" THEN PRINT TAB 28
;(INT (10*T(N)))/10
2150 NEXT N
3000 REM RERUN
3010 PRINT ,,,"FOR RERUN ENTER C
ONT"
3020 STOP
3030 CLS
3040 PRINT "ENTER SECTION NO WIT
H CHANGE"
3050 INPUT N
3060 PRINT "ENTER,DIRECT MODE,QU
ANTITIES/VALUES TO CHANGE, THEN
CONT"
3070 STOP
3060 GOTO 350
4000 REM SAVE
4010 SAVE "TRAP AN█"
4020 GOTO 10
```

fig. 2. Trap antenna program for ZX-81. (For copy of program written for Commodore 64, send SASE to *ham radio*.)

coax can be used for traps

There is a special feature used in the program. This is a routine to calculate the number of turns of RG-58U coaxial cable required to resonate as a trap, as described by Johns, W3JIP,[2] and to calculate the inductance. If other coil diameters are needed, some experimental coils should be constructed and their frequency measured. The number of turns required:

$N = 135.7f^{-0.91}$ on 7/8 inch form
$N = 68.86f^{-0.86}$ on 1.5 inch form

The inductance is calculated from Wheeler's formula,

$$L = \frac{(D + 0.2)^2 N^2}{18 (D + 0.2) + 8N}$$

where the 0.2 is the diameter of RG-58. These relations must all be changed if coax of another diameter and capacity per foot is used.

The final step in the program is conversion of the dimensions and printing the results. The sketch of **fig. 1** should aid in keeping the quantities organized. (In the above equations, a prime is equivalent to changing a subscript).

program listing

The BASIC program for the calculation is given in **fig. 2**. Each section is set off by an REM statement to correspond to the program outline. The program is in Sinclair BASIC, but should run on any common home computer with minor changes. (A version for the Commodore 64 is available from *ham radio*; send SASE.)

```
FREQ DIA LUH CPF HDG  HIN   TRNS
28.5 1   0.6 46  90   94.3  6.4
21.2 1   0.9 60  16.5 26.2  6.4
14.2 1   1.4 86  22.3 47.3  12.1
7.1  1   2.9 169 36.3 155.1 22.7
3.8  1   0   0   35.2 282.2 0

FOR RERUN ENTER CONT
```

fig. 3. Sample run of ZX-81 program for five-band trap antenna using 7/8 inch coax coils.

Fig. 3 shows a sample run. A five-band version using only one-inch tubing would be difficult to build, mechanically, but could be done by using glass-fiber lines for support. A three-band version using two traps is very practical. The writer, as experimental station KK2XJM, uses a six-band version covering 10, 12, 15, 18, 20, and 30 meters.

The main calculation routine is also available for use on the HP 67/97/41 series of calculators.[3]

construction hints

Note that these programs can be used for dipoles by entering wire diameter and by considering that the base section is measured from the center of the dipole to the first trap. When designing antennas using these programs, it is usually best to set the capacitance, since it is most difficult to change. Typical values for dipoles would be 25, 35, 50, and 100 pF for the 10-40 meter traps.

The *ARRL Antenna Handbook* has some hints on construction, and there have been many articles on construction of both integral traps for beams and discrete ones for dipoles. Lately, the author has used only the coaxial trap design for HF. Trap tuning in all cases is by changing coil turn spacing. Sections should be built to allow some length adjustment — about four inches. If necessary, sections can be cut shorter, but the need for this should be rare.

To avoid making up new sections if they must be lengthened, a form of "capacity hat" may be used. This can be two lengths of small tubing clamped to the section to be lengthened at 90 degrees to each other. Tuning is accomplished by moving the "hat" toward or away from the top end of the section. The total length of each added section of tubing should be about twice the added length of section needed.

If the section lengths are excessive, the required lengths can be reduced by using the "capacity hat," a loading coil, or both. To calculate the effect of a loading coil, introduce a dummy frequency, a few percent lower than the value of the next higher band frequency, and use a trap inductance equal to the value of the loading coil contemplated. It will probably be necessary to make several trials to arrive at reasonable values (see sidebar).

Note that these programs are also usable for resonant single-band antennas because the length of these is equivalent to the length of the first section of the trap antenna. Boyer's articles provide information that allows calculating the SWR versus frequency for these trap antennas.

operate both phone/CW

Most trap verticals commercially available require separate settings in order to operate the phone or CW segments of the 40 and 80-meter bands. The following design provides a low VSWR at both sections of each band. Overall length is just under 50 feet.

FREQ	DIA	L-UH	C-PF	H-DEG	H-IN
7.2	1.5	6.5	75	89.9	376
7.025	1.5	6.8	75	1.3	5
3.795	1.5	17.5	100	23.9	191
3.525	1.5	0	0	2.9	25

A more manageable height (38.8 feet) is achieved by using a loading coil for the 3.5 MHz portion:

FREQ	DIA	L-UH	C-PF	H-DEG	H-IN
7.2	1.5	6.5	75	89.9	376
7.025	1.5	6.8	75	1.3	5
6.75	1.5	55.5	10	1.5	7
3.795	1.5	17.5	100	7.8	62
3.525	1.5	0	0	1.4	12

A small further reduction in height can be obtained by changing the section above the last trap to a top hat: the diameter should be about one-half the length of the section. If still further reduction in height is needed, another loading coil could be placed near the top of the 40-meter section.

The trap quality in these designs must be very good. The coil should be about twice the diameter of the section, made of large wire or small tubing, and spaced to give a length about equal to the diameter. The 7.025 MHz trap and the loading coil can be wound as one coil, tapped for the capacitor connection.

These designs have not been tested. Because of the closeness of trap frequencies, expect to do some adjustment after construction.

references

1. Joseph M. Boyer, W6UYH, "The Multi-Band Trap Antenna," *CQ*, February, March, April, May, 1977, pages 26, 51, 46, 22 respectively.
2. Robert H. Johns, W3JIP, "Coaxial Cable Antenna Traps," *QST*, May, 1981, page 15.
3. R.P. Haviland, "Trap Antenna," *HP Library Program 97-04766-2*, Corvallis, Oregon.

ham radio

the "smart" frequency counter

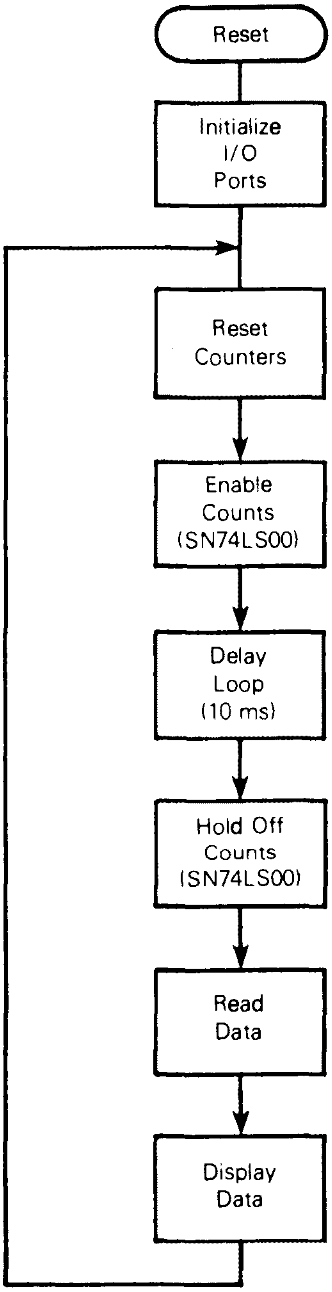

fig. 1. Flowchart for basic frequency counter.

Build a versatile test instrument with digital display and microprocessor control — for less than $50

Digital displays for radios have been around for several years, and most people have grown quite used to their appearance and operation. After purchasing an inexpensive HF transceiver without a digital display, I actually felt a little lost because I had been used to seeing the digits flicker as I tuned across the band. So I decided to add a digital display to my unit that would also function as a normal frequency counter.

This article describes not only how to build a general-purpose microprocessor-controlled frequency counter, but also provides some routines that will allow the counter to be used as a digital display for a radio. In production quantities, this frequency counter could be built for about $10.00 or $15.00. In single quantities, a price of less than $40.00 to $50.00 would be more realistic. A parts list at the end of the article provides prices and quantities required.

In trying to determine the actual operating frequency of a radio, several problems can arise. Depending on the type of radio (single conversion, dual conversion, direct conversion, etc.), different counting methods must be used to determine operating frequency.

It's easiest to provide a digital display for a *direct conversion* type because the actual VFO frequency is the operating frequency. In fact, any frequency counter may be used to measure the operating frequency.

On *single conversion* radios, the frequency to be counted is either the sum or difference of the VFO and HFO. In this configuration, several things must be considered: if selectable sideband operation is required,

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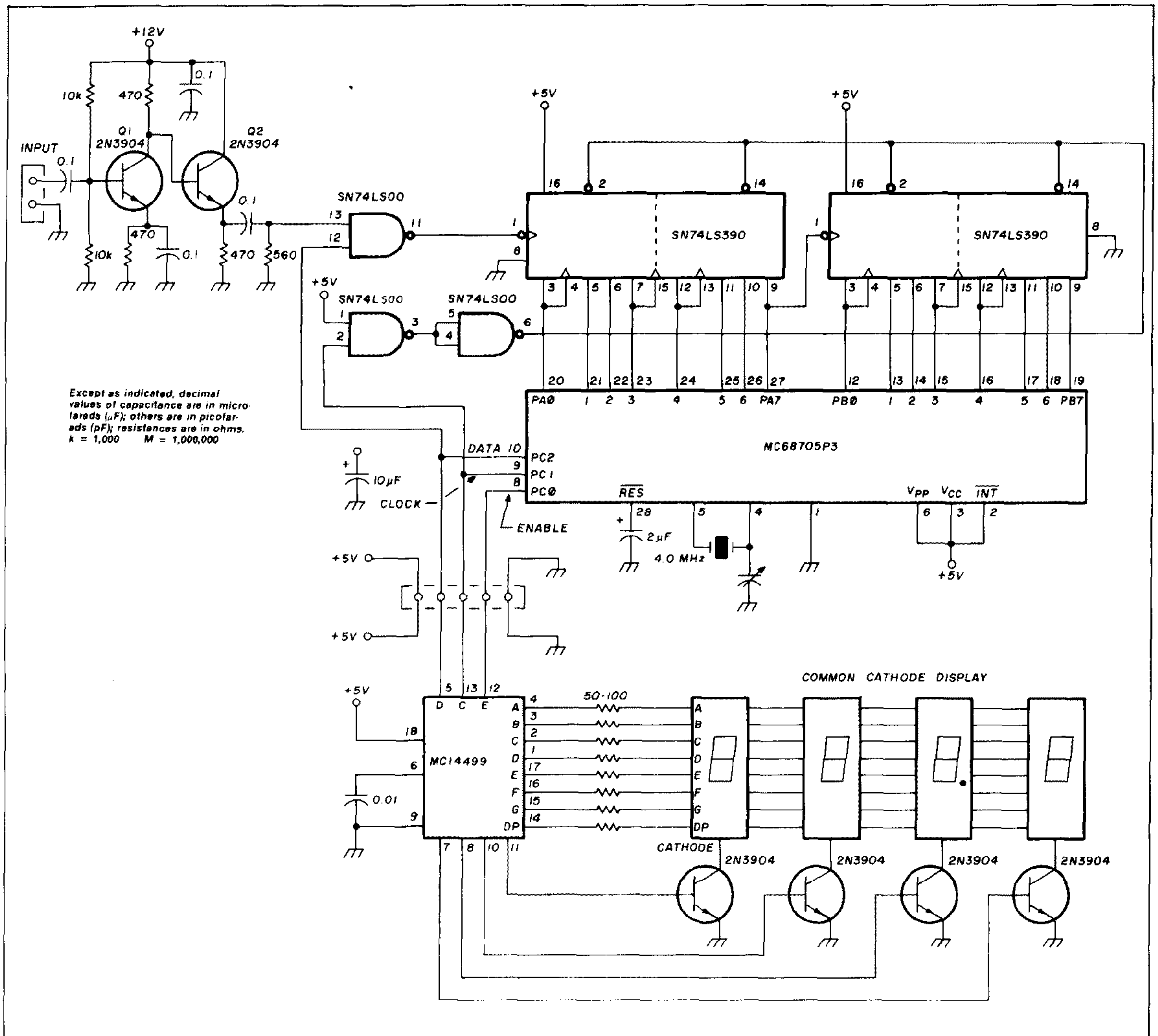


fig. 2. Basic frequency counter schematic.

some provision must be incorporated to allow shifting of the displayed frequency. Generally, this is an additional wafer on the mode switch. Also, in counting the frequency, the MHz may or may not be valid (with respect to the operating frequency), but the 100 kHz (and down) digits would be correct. To provide a correct MHz reading, an additional wafer on the bandswitch is required. One alternative is to read the HFO, read the BFO, and either add or subtract the two, depending upon which is necessary.

On *double conversion* types of radios, the counter must be preset to get the correct frequency, and the mode taken into account. On both the single and double conversion radios, it is obvious that some necessary "smarts" must be incorporated by the counter to make it display the proper frequency. A normal frequency counter could not easily be modified to provide these features; instead, a microprocessor would be an ideal choice for this function.

description

The heart of the counter is the MC68705P3 microprocessor, an EPROM, which means that a user can erase the internal program using ultraviolet light, and reprogram it many times. (In a production-type environment, a mask-programmed part would be used. Mask programming is done as a one-time shot at the factory, and is much less expensive than the EPROM part.)

Programming the EPROM is a simple task that may

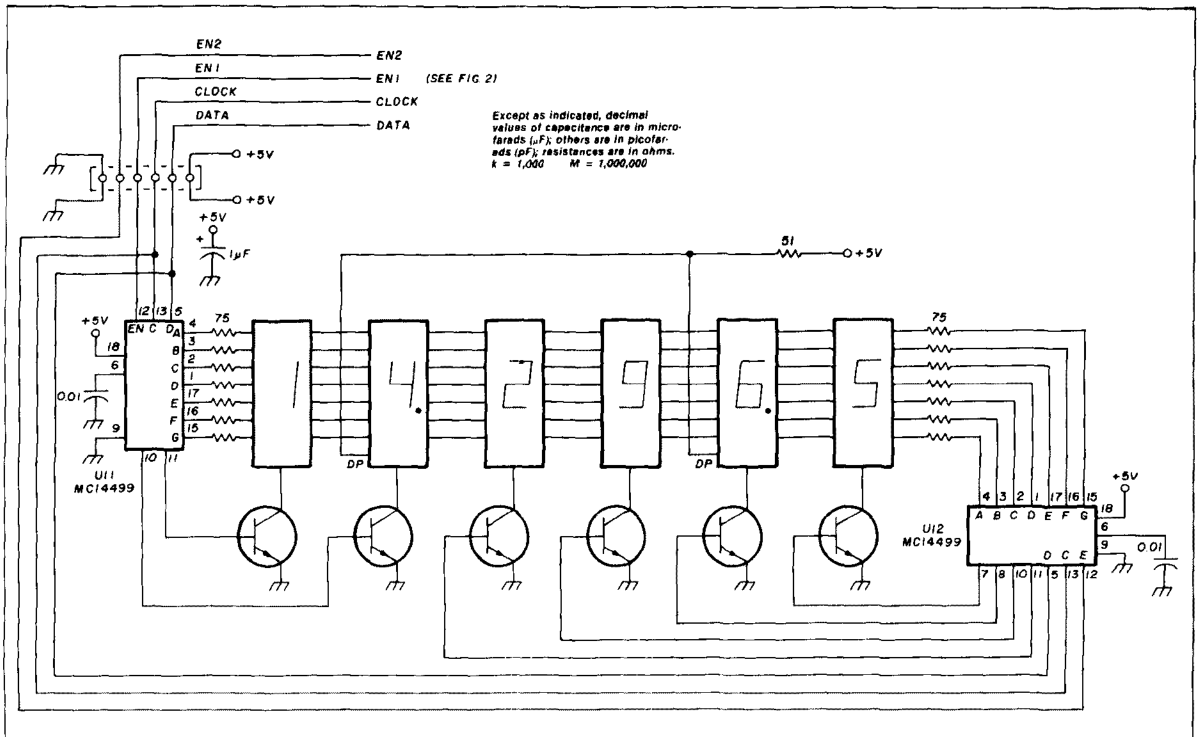


fig. 3. The serial display can be separated from the microprocessor.

be done by anyone; *Motorola's Application Note AN-587* and the accompanying data sheet contain instructions.

operation

Most frequency counters use a high frequency time base, divide it down, and use it to allow a certain number of input pulses to go into some BCD counters. These counters then feed the normal latches and BCD to 7-segment decoder drivers for the display. In the MCU (microcomputer unit) version of the frequency counter, the MCU controls the actual time base generation, latching of data, and driving of an external display. A basic block diagram of the system is shown in **fig. 1**.

The inputted analog signal is conditioned and amplified, which means it's turned into a digital signal that the computer system can use. A two-input NAND gate is fed by both this signal and another generated by the computer. The computer-generated signal actually inhibits the input signal from propagating through the counters. Only by having the capability of turning off the frequency source can an accurate count be made. The output of the gate (NAND) is fed into the BCD counters, which do the actual counting. These counters have a "clear" input on them which allows them to start at a count of 0000. This signal is, of course, controlled by the MCU. After being cleared, the LS390s count until the 10 millisecond period (generated by the MCU) has been completed. At this time, the MCU disables the gate, reads the data in the counters, and sends that information serially to the display drivers.

The most critical component of any frequency counter is the time base, and this device is no different. A 4 MHz crystal is used as the clock for the MCU, which in turn controls the actual 10 millisecond gate period. To adjust the counter, the highest frequency to be counted should be measured by an external frequency counter, and the trimmer capacitor on the crystal should be adjusted to give an identical reading. For further stability, an external temperature compensated oscillator may be used. Do not try to count the time base by measuring one of the crystal pins. Even an extra load of 5 pF is sufficient to shift the time base frequency and degrade the accuracy of the counter.

The basic frequency counter schematic (**fig. 2**) can be easily added to a radio such as the Ten-Tec Argosy. Note that while the MHz display option may be incorporated through the addition of a bandswitch, further cost reductions may be realized by only having four digits displayed, with the MHz being read off the band switch.

reducing computer-generated noise

In any system that uses a computer, some noise is

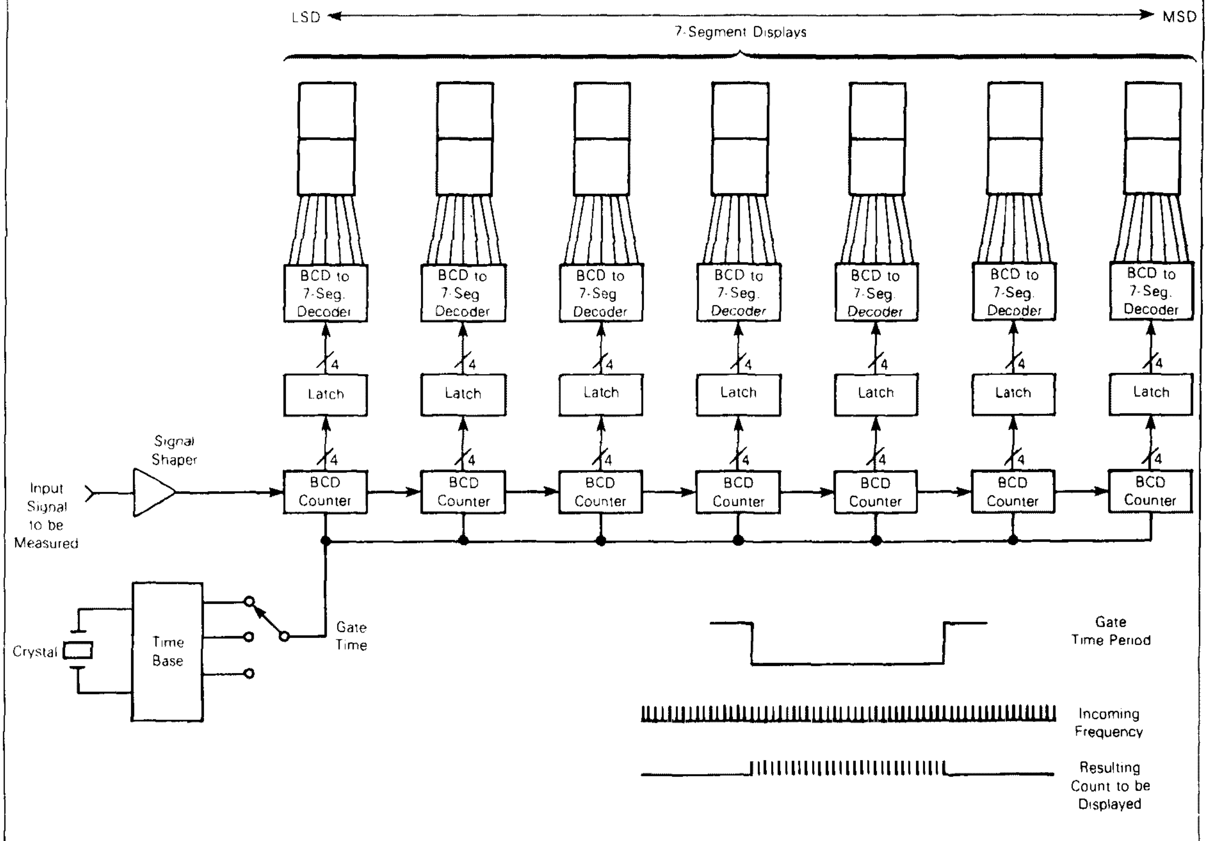


fig. 4. Basic frequency counter block diagram.

generated. Even with an MCU, which has most of the noisy circuitry inside the ceramic package, fast transitions of the I/O lines will cause some RFI. I recommend that any computer used in a radio environment be shielded. This may be done in several ways. One method employs thin PC board or copper flashing to make a box that totally shields the MCU. This cuts down on most noise that may be generated from the computer. This is especially necessary when the computer is of the "wire-wrap" or hand-wire construction.

expandability in a single chip MCU

With the advent of single-chip microcomputers, many new "peripheral" parts (those which do some function external to that of the main computer chip) are being designed to "talk" serially. By sending the data down lines in a serial bit stream, many extra I/O pins may be saved for other functions. That is the name of the game in an MCU environment, because once all I/O lines have been used, sometimes a total re-design of the system is required to get "just one more." A good example of this concept of serial peripherals is shown in the schematic of **fig. 2**.

The display device used, the MC14499, is of this serial variety (saving I/O pins). Each device is capable of driving a total of four seven-segment LEDs, but the first counter example requires only one. In more sophisticated systems, two 499s could be used to provide a total of six digits. The MC14499 is a simple part to use, and the only external components required are eight current limiting resistors (seven for the segments, and one for the decimal point), four transistors, and a small capacitor. This makes for an easy layout, and since the data is sent serially, the display can be remote from the actual MCU. **Fig. 3** shows the display schematic and required hardware.

To re-emphasize the importance of serial-type peripheral devices, consider the number of I/O lines which are required to drive four digits of LEDs. To do this directly, at least sixteen lines are required. This would really be a waste! By using the serial format (data, clock, and enable), only three lines are needed for the first display driver. To add additional drivers, add only one more enable line for each device. The data and clock lines are common.

alternate displays

If an LCD type of display is desired, an MC145000

```
                    *
                          NAM    CNTR
                          OPT    PAG
                    *
0000                PORTA EQU    $000
0001                PORTB EQU    $001
0002                PORTC EQU    $002
                    *
0004                PADDR EQU    $004
0005                PBDDR EQU    $005
0006                PCDDR EQU    $006
                    *
0040                SAVEA EQU    $040
0041                SAVEB EQU    $041
0042                TEST  EQU    $042
                    *
0784                      ORG    $784
                    *
0784 07             MOR   FCB    $07      CRYSTAL MODE
                    *
                    * BEGINNING OF CODE
                    *
0080                      ORG    $80
                    *
0080 3F 04          START CLR    PADDR    ALL INPUTS
0082 3F 05                CLR    PBDDR    HERE TOO
0084 A6 0F                LDA    #$0F     ALL OUTS
0086 B7 06                STA    PCDDR
                    *
0088 12 02          CONTIN BSET  1,PORTC  CLEAR OUT COUNTERS
008A 13 02                BCLR   1,PORTC
                    *
008C 14 02                BSET   2,PORTC  ENABLE COUNTER
                    *
                    * BEGINNING OF TIMING LOOP
                    *
008E A6 ED                LDA    #$E0
0090 9D             LOOP  NOP
0091 9D                   NOP
0092 3A 42                DEC    TEST
0094 3A 42                DEC    TEST
0096 3A 42                DEC    TEST
0098 3C 42                INC    TEST
009A 3C 42                INC    TEST
009C 4A                   DECA
009D 26 F1                BNE    LOOP
                    *
                    * A LITTLE MORE FINE TUNING
                    *
009F 3C 42                INC    TEST
00A1 3C 42                INC    TEST
00A3 3C 42                INC    TEST
00A5 3C 42                INC    TEST
00A7 3C 42                INC    TEST
00A9 9D                   NOP
                    *
                    * FINALLY, ALL COUNTED
                    *
00AA 15 02                BCLR   2,PORTC  HOLD OFF ANY MORE COUNTS
                    *
                    * NOW WAITED 10 MS
                    *
00AC B6 00                LDA    PORTA    GET LS DIGITS
00AE B7 40                STA    SAVEA
                    *
00B0 B6 01                LDA    PORTB    GET MS DIITS
00B2 B7 41                STA    SAVEB
                    *
                    * SEND OUT DATA TO DISPLAYS
                    *
00B4 11 02          SNDDTA BCLR  0,PORTC  ENABLE = 0
00B6 12 02                BSET   1,PORTC  CLOCK = 1
                    *
00B8 AE 05                LDX    #$05     NUMBER TO ROTATE
00BA A6 20                LDA    #$20     DECIMAL POINT HERE
00BC AD 0E                BSR    CLOCK1
                    *
00BE B6 41                LDA    SAVEB    GET LSB FIRST
00C0 AD 08                BSR    CLOCK
                    *
00C2 B6 40                LDA    SAVEA    NOW MSB
00C4 AD 04                BSR    CLOCK
                    *
00C6 10 02                BSET   0,PORTC  NOW ENABLE IT
                    *
00C8 20 BE                BRA    CONTIN   CONTINUE COUNTING
                    *
                    *
00CA AE 08          CLOCK LDX    #$08
00CC 49             CLOCK1 ROLA
00CD 24 0A                BCC    ZERO
00CF 14 02                BSET   2,PORTC
00D1 13 02          CLOCK2 BCLR  1,PORTC  SHIFT IN ONE BIT
00D3 12 02                BSET   1,PORTC
00D5 5A                   DECX
00D6 26 F4                BNE    CLOCK1
00D8 81                   RTS
                    *
00D9 15 02          ZERO  BCLR   2,PORTC  MAKE A ZERO
00DB 20 F4                BRA    CLOCK2   GO CLOCK IT
                    *
                    * RESTART VECTORS
                    *
07F8                      ORG    $7F8
                    *
07F8 00 80          TIMER  FDB   START
07FA 00 80          EXTINT FDB   START
07FC 00 80          SWIINT FDB   START
07FE 00 80          RESET  FDB   START
                    *
                          END
```

fig. 5. Software for the MC68705P3.

may be used to drive a six-digit multiplexed LCD. The data format is a bit different than that of the 14499, but it is also a serial device and quite easy to use.

Several other serial parts — and microcomputers to talk to them — are becoming available. The MC144110 is a 6-bit D to A converter; the MC144102 is a 16 × 16 bit RAM; and the MC145157, 58, and 59 are serially-controlled PLLs. Look for more functions to be available in the future which use this type of data transfer.

The ability to use these serial devices is greatly enhanced by the instruction set of the M6805 family of processors. Through the use of the "Bit Set/Bit Clear" instructions, any I/O pin or RAM bit may be set or cleared with a single instruction. (This is in contrast to other, generally older, microprocessors that need to get the entire 8-bit port into a single register, and then do an appropriate Boolean instruction to set or clear the bit — AND, OR, XOR, etc. The entire word then had to be stored back to that port. This required several instructions and a lot of time to implement.) With true bit-manipulation, any type of serial data transfer is easy; several other instructions also simplify work in a controller environment. The "Branch if bit Set/Branch if bit clear" instruction uses this bit manipulation architecture in a branching situation to do things that depend on what an individual bit's state is.

software

A flow chart of the "simple" 4-digit counter is shown in **fig. 4**, and the actual software for the MC68705P3 is shown in **fig. 5**.

future developments

From the discussion above, it can be seen that increasing the range of the counter is quite simple. Additional counters, prescalers, or other range-extending devices may be implemented. By placing these prescalers in front of the 74LS000 gate, frequencies greater than 1500 MHz may be counted. Or how about a full-function DVM & frequency counter all in one small package? By using the MC68705R3, which also incorporates an 8-bit A to D system on chip, a complete DVM could be implemented with appropriate scaling resistors.

ham radio

the ham notebook

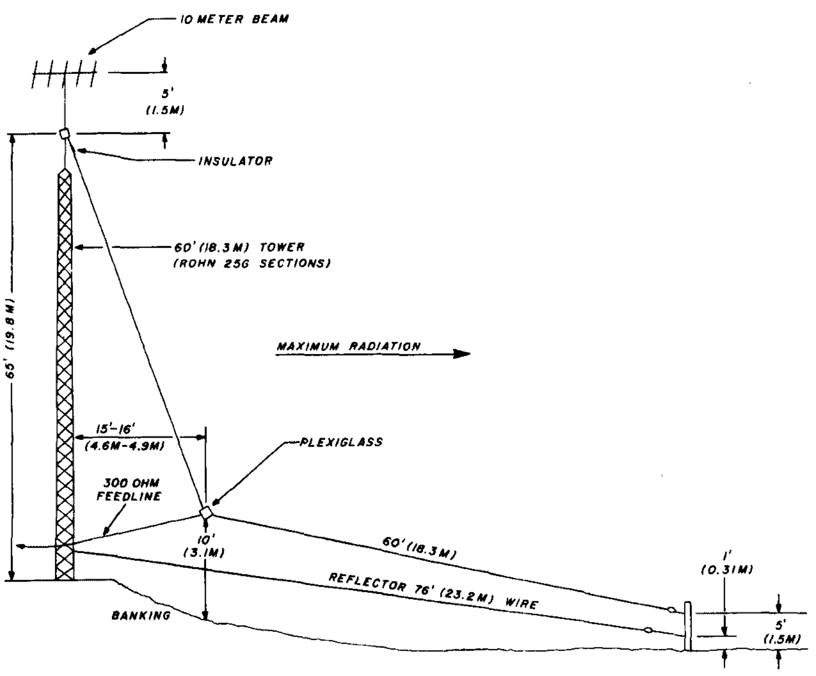


fig. 1. Two element 80 λ parasitic array produces low-angle radiation. (The tower is guyed with Dacron rope. If wire is used, insulate every 15 to 20 feet with egg strain insulators.)

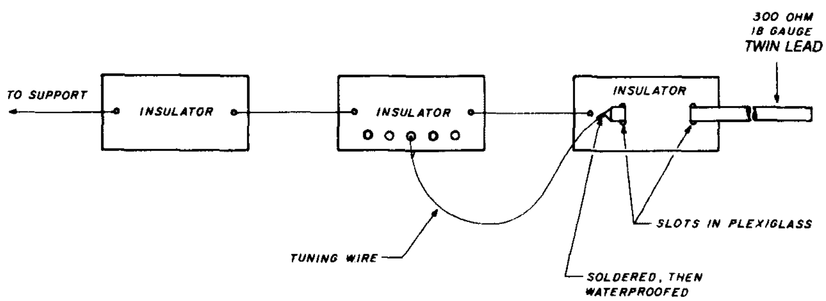


fig. 2. Fine length adjustment for resonance uses single tuning wire at antenna ends.

80-meter half-wave sloper uses reflector

This antenna was the result of a desire to work 80-meter DX running low power. Many excellent reports have been received, including long path to VK, ZL, and JA.

The antenna, basically a dog-legged, 1/2-wave sloper, uses a full-length reflector (**fig. 1**). The reflector is the tower plus enough wire hooked to the base and extended out under the sloper to give a full-length (1/2 plus 10 percent) reflector. The driven element is a folded dipole which can be fed with 300-ohm feedline or used 50-ohm line when a 4:1 balun is inserted (I used the 300-ohm feedline). Because the amount of doglegging affects the feed impedance, it should be adjusted for minimum SWR. (The closer the antenna is to the tower, the lower the impedance). I used 18 gauge 300-ohm ribbon for QRP; later, when I ran 800 watts, no heating was noted. 18 gauge can easily handle 2 amperes, or 1200 watts at 300 ohms; open wire 300-ohm line is a better choice if available. The dog-legging causes the pattern to "squint" in the direction of the dipole. I estimate the gain (over a dipole) to be 4 to 6 dB. The angle is low enough to produce good long path results. (The author has been able to consistently contact VK6LK-Robin almost throughout the entire year — Ed.)

Use very good end insulators for the driven element to prevent arc-over or leakage. Plexiglas (1/4-inch thick) works very well and is easy to obtain as scraps from manufacturers or others working with it. With the feedline connected to the physical center of the antenna, the portion of the dipole toward the tower may have to be shortened 6 to 15 inches in order to

maintain electrical symmetry. The end of the folded dipole is connected to a single wire to facilitate tuning (see **fig. 2**). Last but not least, use a rope and pulley to connect it to the top of your tower unless you really like to climb.

Bruce Clark, KO1F

Argonaut 509 conversion for 30 meters

As declining HF propagation renders the 10-meter band less of a dependable mainstay for the QRP operator, the new allocation at 10 MHz is coming into its own for this mode of activity. The operating restrictions imposed upon 30 meters actually help rather than hinder QRP operation.

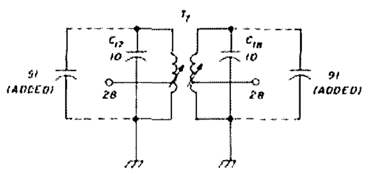


fig. 1. Modification of the Argonaut Model 509 transmitter section only requires adding two capacitors to the T7 tuned circuit on the 80262 board.

The Ten-Tec Argonaut 509 may easily be modified to cover 30 meters with the same performance found on its five bands. All that is required for the modification is five minutes of time, four components, and a jumper wire.

Ten-Tec uses a 9-MHz IF with appropriate VFO frequencies. On 10 meters, the VFO ranges from 19 to 21 MHz. If we use the difference mixer product, rather than the sum, as in the original design, output occurs on 10 MHz rather than 28 MHz. Since the 509's transmitter stages are broadband amplifiers, the only changes required for 10 MHz transmit are to re-resonate the appropriate bandpass filter, composed of T7 and capacitors C17 and C18 on the 80262 front-end board (see **fig. 1**). Shunt C17 and C18 each with an additional 91 pF of capacitance. Depending upon component tolerances, retweaking T7 cores may be necessary for adequate output. The addition of the two 91-pF capacitors completes transmitter modification.

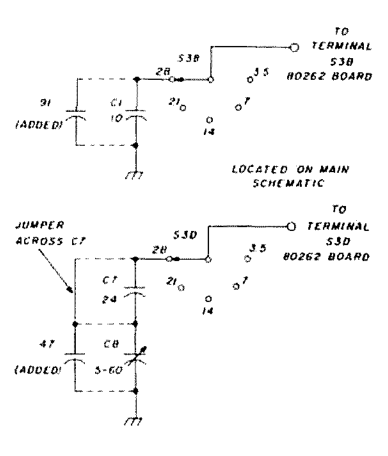


fig. 2. Modification of the receiver section of the Argonaut 509 requires minor additions to components switched by S3B and S3D.

The receiver will function with much reduced performance without modification by turning the receive preselector completely counterclockwise, as mentioned in the 509 manual for WWV reception. To improve 10 MHz performance the receiver front end needs additional capacitance.

C1, on the main schematic, is switched across T1 to resonate at the desired band. For our purposes, C1 needs to be shunted with another capacitor, again of 91 pF, as seen in **fig. 2**. T2, on the other end of the RF amplifier front end, must also be resonated at 10 MHz. Shorting C7 with a piece of tinned bus wire and parallelling variable C8 with an additional 47 pF completes our modification.

The modification is simple and quick. Lifting one end of each added component easily restores 10-meter operation when the sunspot activity increases.

Modification of the newer Model 515 is identical. The rig that began it all, the 505, is not broadbanded on transmit.

Raymond Henry, AA4LL

audio oscillator to pulse generator conversion

Here is a quick, simple method for converting an audio oscillator into a pulse generator that has served well on bench digital projects. (See **fig. 1.**)

To cover the full range from 10 Hz to 1 MHz, C1 and R1 may be either fixed in values with C1 from 0.1 to 0.001 μF and R1 from 10 k to 1 megohm, or C1 fixed at 0.1 and a 1 megohm pot used for R1. The drive input from the oscillator should be 1.5 to 3.5-volts rms if Q1 is a 2N2222 and a 0.75-volt rms if Q1 is an MPF6515.

The pulse width is directly related to the oscillator frequency, with the narrowest being 0.6 μs. For applications requiring fast pulse triggers, the similar circuit using Q2 and the unused two sections of IC1 will provide a fine 1 μs pulse with polarity selected from the IC1 outputs as shown.

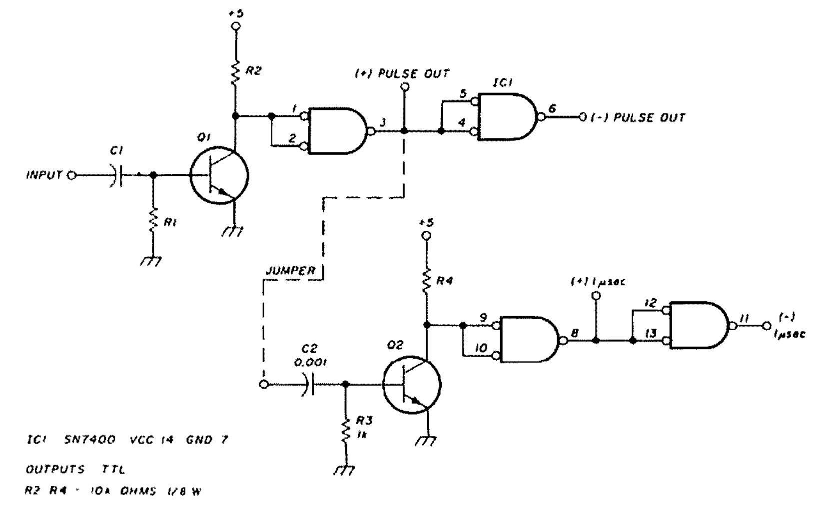


fig. 1. Audio oscillator conversion schematic.

Assembly can be accomplished on a piece of perf board with direct pin wiring in about 30 minutes.

Gene Shapiro, WØDLQ

VHF/UHF WORLD

Joe Reisert W1JR

VHF/UHF frequency calibration

Have you ever wondered whether you're really on frequency — especially when you don't hear the station you're scheduling? This is a constant problem for the VHF/UHFer, especially during meteor scatter or EME schedules. Because the focus of this issue is on test equipment, I thought it would be a good time to discuss frequency accuracy and offer some advice on accurately determining frequency. Then I'll describe a secondary frequency standard or calibrator you can build to satisfy most of the needs *and* pretty much guarantee that you're at least on the frequency you think you're on!

frequency determination

There are several ways to determine frequency. You can:

- read the frequency indicator on the gear in use;
- trust the frequency of the crystal in your up or down converter;
- have a reliable friend measure your transmitted frequency;
- buy or build a good frequency counter; or
- build or buy a good secondary frequency reference standard.

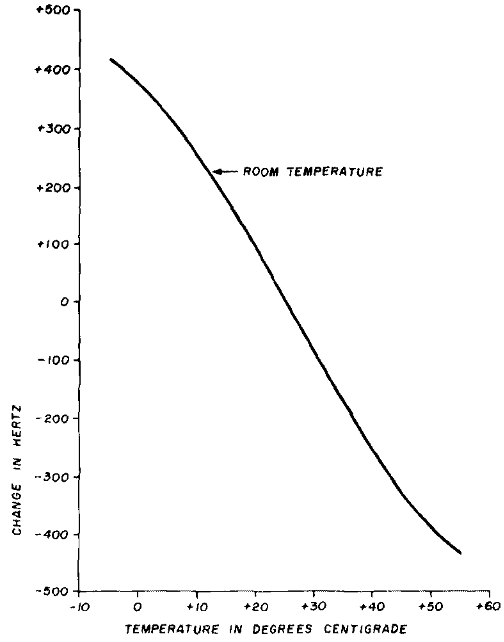


fig. 1. Typical frequency drift versus temperature for a 48-MHz overtone oscillator.

All of the above methods have advantages and disadvantages. While you may laugh at the thought of just reading your dial or frequency indicator, some modern commercial gear transmits CW at an offset frequency that may not be directly indicated.

Recently I tested one of the latest state-of-the-art VHF/UHF multi-mode transceivers complete with a built-in digital frequency readout. To my amazement, the frequency indicated was off by almost 1 kHz in the first 15 minutes; even after an hour the frequency *indicated on UHF* was off by over 0.5 kHz — admittedly not a serious problem, but one worth examining if frequency accuracy would affect the success of a QSO or the accuracy of measuring someone else's transmitted frequency.

crystal oscillators are used everywhere

All modern converters or transverters use crystal oscillators in one form or other. However, if 1.0 kHz accuracy is required at UHF, the cost of a crystal is prohibitive. Furthermore, the accuracy of the oscillator is also a function of the oscillator circuit and the temperature of the crystal. Some designs use small trimmers to tweak the frequency to the marking on the crystal, but this can have disastrous effects on phase noise, not to mention decreased temperature stability or drift.[1]

All crystals, regardless of price, suffer from a problem called "aging." This means that a crystal will always be drifting slightly away from its marked frequency even after it has been burned in for a predetermined time. (The typical commercial standard is full power for 30 days.)

Suppose a friend measured your

transmitted frequency. Did he or she tune you in properly on SSB? on CW? Was he or she at true zero-beat? Even if the measurement is accurate, how do you account for local ambient temperature changes, aging or drift and component changes in your own gear at a later time and date? How long does it take to properly warm up your gear?

Frequency counters are great, but until recently they were beyond the reach of most Amateurs. Several are now available, but suitable ones use a reference standard and a proportional control oven that takes time to stabilize — perhaps 15-30 minutes — and usually cost about as much as a typical multi-mode transceiver! Furthermore, frequency counters that read higher than 600 MHz are even more expensive.

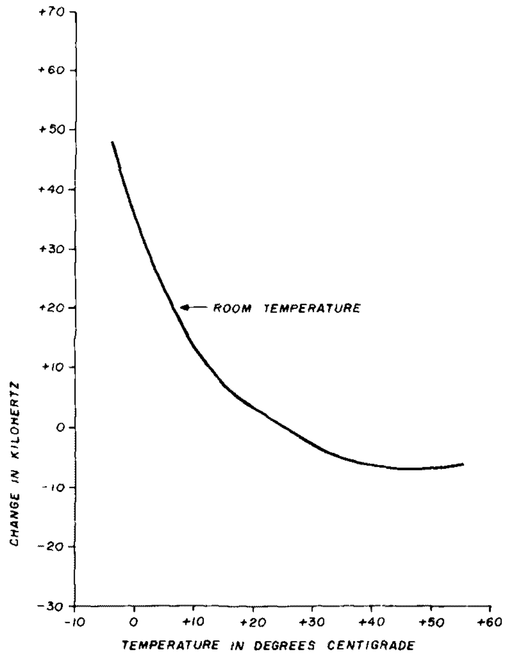


fig. 2. Typical frequency drift versus temperature for a 1.000 MHz secondary frequency standard.

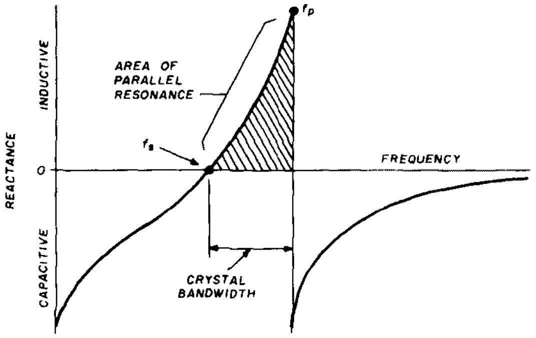


fig. 3. Typical impedance versus frequency variation for a quartz crystal operating in the fundamental mode. "f_s" is the series resonant frequency and "f_p" is the parallel or antiresonant frequency. The typical bandwidth is 5-10 kHz at 4.000 MHz. The parallel resonant frequency can be varied by the use of an external capacitor in series or parallel with the crystal.

there is another way

A good method that seems to have been almost forgotten by the users of modern rigs is the secondary frequency standard or crystal calibrator. If one is available with a comb generator (a device which generates many harmonics of the input frequency), it can be used up through UHF with surprising accuracy.[2]

After evaluating all these alternatives to accurate frequency determination at VHF/UHF frequencies, I decided many years ago that the least expensive and most accurate way to measure my frequency was to build a secondary frequency standard. Before designing one, I listed the features I wanted to include:

- convenient marker frequency in each VHF/UHF band above 6 meters
- good short and long-term stability, 1 to 2 parts per hundred million (0.01-0.02 ppm)
- easy adjustability to the correct frequency
- high harmonic output through 2304 MHz
- battery operation
- fast warmup
- reasonable size, including built-in antenna
- easy construction
- low cost

This is indeed quite a "wish list!" However, after some trial-and-error (over a period of 15 years) such a unit has evolved.

The most important consideration for a secondary standard is crystal stability. First, I tried a third-overtone 48-MHz crystal oscillator. Harmonics were no problem, but stability, the very property I wanted to measure,

was no better than the local oscillators in my receiver or transmitter — about 3 to 4 parts in 10 million per degree Centigrade (fig. 1). Also, the only usable harmonics were on 144, 432, and 1296 MHz.

Next, a 1.000-MHz calibrator was tried, but proved quite unsatisfactory for several reasons.[2] First, it is difficult to produce sufficient harmonics to reach the lower UHF bands. After all, for 70 cm (432 MHz), that is the 432nd harmonic! Second, and more important, the frequency stability of a good 1-MHz crystal versus temperature is quite poor. I measured such a unit at typically 500 to 600 parts per million per degree Centigrade (see fig. 2).

After discussing this with crystal manufacturers I discovered that reasonably priced, stable crystals and oscillators can be designed and built without an oven if you choose an "AT" cut fundamental crystal in the 3 to 10 MHz region (more on this later). 4 MHz was chosen because harmonics would be present on all of the VHF bands above 6 meters, lending itself to easy calibration with WWV on a 20-MHz HF receiver. Later I built a 5-MHz unit to be used on 6 meters and other multiples of 5 MHz.

the circuit you choose is important

Crystals are usually classified as either fundamental or overtone. Because of cost and stability constraints, fundamental crystals are usually used below and overtones above 22 MHz. Each fundamental crystal has two basic resonances, a series and a parallel or anti-resonance. The series resonant frequency is typically a few kHz below the parallel resonant frequency and is very difficult to externally move or adjust. However, the parallel mode resonant frequency can be easily adjusted by placing a capacitor either in parallel or series with the crystal. These characteristics are shown in fig. 3. A parallel mode oscillator was chosen for this reason. The Pierce oscillator circuit was first used because it is reputed to be the most stable, but the parasitics of the active device in the oscillator are prominent. Therefore, a modified Colpitts oscillator circuit was later chosen because it was only slightly less stable than the Pierce oscillator, but more easily adjustable, and uses swamping capacitors to minimize the parasitics of the active device.

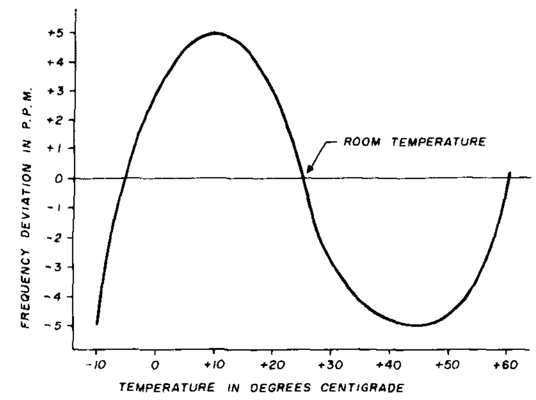


fig. 4. Typical frequency drift versus temperature for an "AT" cut high-accuracy crystal at 4.000 MHz.

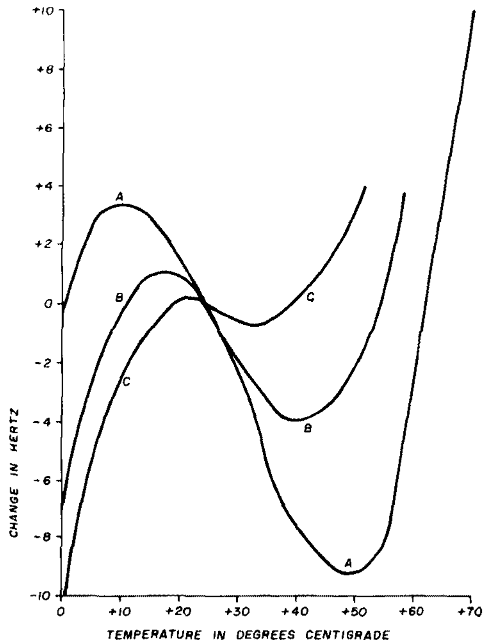
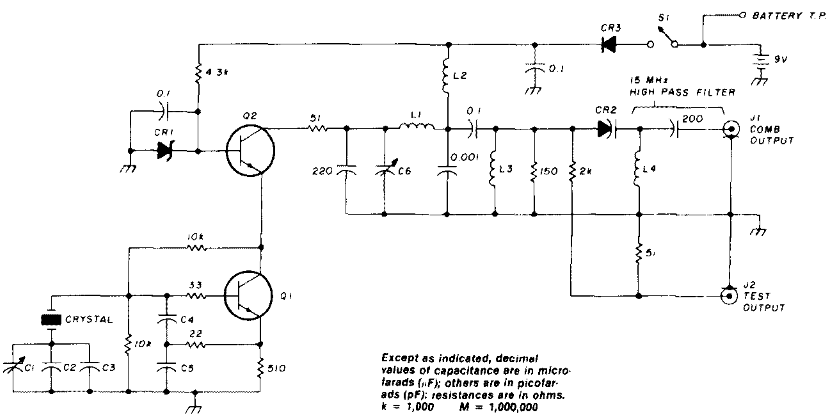


fig. 5. Frequency versus temperature variation of the 4.000 MHz calibrator in fig. 6. For various amounts of temperature compensation "A" is undercompensated while "B" and "C" are optimum versus desired temperature range, toward either cold or hot.

"AT" cut crystals have a known frequency versus temperature characteristic that is "S" shaped (fig. 4). Hoff has pointed out that if you want to operate only over a small temperature range such as experienced in most ham shacks, the frequency change is



item	description
C1	1-10 pF air variable capacitor, Johanson 5200 series or equivalent
C2,C3	10 pF N750 ceramic disc capacitor (see text)
C4,C5	390 pF NPO ceramic disc capacitor (see text)
C6	5-50 pF small ceramic or mylar variable capacitor
CR1	4.7 volt zener diode, 1N750 or equivalent
CR2	SRD, Microwave Associates No. MA44310, Hewlett-Packard No. 5082-0180, or equivalent (see text)
CR3	light emitting diode
J1,J2	BNC chassis mounted connector
L1	6.8 μH RF choke
L2,L3	27 μH RF choke
L4	0.47 μH RF choke
Q1,Q2	2N2222 transistor
S1	small toggle switch
crystal	4.000 MHz parallel mode 20 pF crystal (see text)

fig. 6. Schematic diagram for a 4.000 MHz secondary frequency standard.

quite linear and can be decreased over a small temperature range (perhaps 10 to 40 degrees Centigrade) by simply placing a temperature compensating capacitor in series or parallel with the crystal.[3]

Indeed, this is what I did. First an NPO (which has close to zero temperature drift) temperature compensating ceramic disc capacitor (in parallel with a tweaking capacitor) was placed in series with the crystal in an oscillator. Then the frequency versus temperature was measured in a laboratory oven in 5-degree Centigrade steps and the frequency was remotely measured with a good frequency counter having a proportional control oven. The NPO capacitor was then replaced with various combinations of 10 pF N750 type temperature compensating ceramic disc capacitors and data taken again. It soon became obvious that the temperature drift could be minimized by the right choice of temperature compensation. This is illustrated in **fig. 5** for a typical 4.000-MHz crystal oscillator.

now for those harmonics

Now that I had the oscillator working, I tried various schemes to generate a comb of harmonics. Integrated circuits were discarded because they required considerable current and did not extend high enough in frequency. Tunnel diodes were also tried but did not meet the UHF harmonic requirements.[2]

SRDs (step recovery diodes), properly chosen, fit the desired requirements. Furthermore, if used in the proper circuit, they required low RF drive and could generate power through 2304 MHz. Oscillator-multiplier isolation is required so that frequency stability or calibration is not affected by the multiplier or output circuitry. This was accomplished by using a separate amplifier or buffer stage between the oscillator and multiplier.

values are now chosen

Several solid-state devices and circuit configurations were tried. JFETs and MOSFETs were eventually discarded in favor of bipolar transistors since the variation from unit to unit necessitated additional tweaking for proper bias and compensation. An oscillator-amplifier approach was used in the first few calibrators but eventually replaced by a cascade bipolar transistor circuit that provided the required oscillator to multiplier isolation with low power consumption and could be powered by a 9-volt transistor radio type battery.

A schematic of the final circuit is shown in **fig. 6**. The modified Colpitts oscillator uses a bipolar transistor, Q1. The trimmer, C1, should be a high quality air dielectric variable with a multi-turn fine adjustment screw. C2 and C3 are the temperature compensating disc ceramic capacitors. The typical oscillator required 20 pF N750 capacitors and in the extreme case required 20 pF N1500 compensation. The feedback capacitors, C4 and C5, should be NPO-type ceramic disc capacitors. Do not use silver mica capacitors here; they have a positive temperature coefficient that will increase the drift!

Crystal specification is most important. Beware of bargain basement or computer-type crystals. They may be specified for series resonance or have a poor temperature coefficient, which will make it nearly impossible to tune or compensate the calibrator at the desired frequency. International Crystals* sells a "HA" (high accuracy) crystal that should meet the requirements. The price may be higher than you're used to paying for a crystal, but because you'll probably use this secondary reference standard for many years the crystal will pay for itself many times over. Remember to specify the crystal as *parallel* resonant with 20 pF of capacitance. The temperature tolerance should be ±0.0005 percent from −30 to 60 degrees Centigrade. The calibration tolerance can be ±0.0025 percent. Over-specifying the actual calibration frequency is a waste of money because the trimmer will adjust you precisely to zero-beat.

*International Crystals, P.O. Box 26330, Oklahoma City, Oklahoma 73126.

reducing supply voltage dependency

A recent addition to the circuit was a zener diode in the amplifier stage. This provides a good reference voltage to the oscillator over a wide battery voltage range with very little extra power consumption. A standard 9-volt transistor radio type of battery, preferably the longer life type, is all that is required because the current drain is typically 4 to 5 milliamperes. Another *recent innovation is the* LED (light emitting diode) in series with the battery. This provides a low-power indicator to remind you to turn off the calibrator when not in use. A battery test point is also provided so that you won't have to open the box to measure the battery voltage.

The multiplier circuit also evolved over the years. A simple pi-network matches the amplifier output to the SRD. The SRD chosen should be of the long lifetime type, typically 100 nanoseconds minimum, since 4 MHz is a very low input frequency for many SRD's. Other types of diodes will probably not work well. C6 peaks the circuit, which matches the amplifier to the SRD. The output of the SRD passes through a 15 MHz high-pass filter, suppressing the fundamental frequency of the comb output. Therefore a calibrator output circuit was added in case you desire to measure frequency directly (such as with a frequency counter) at 4.000 MHz. The typical comb generator output power versus frequency curve is shown in **fig. 7**.

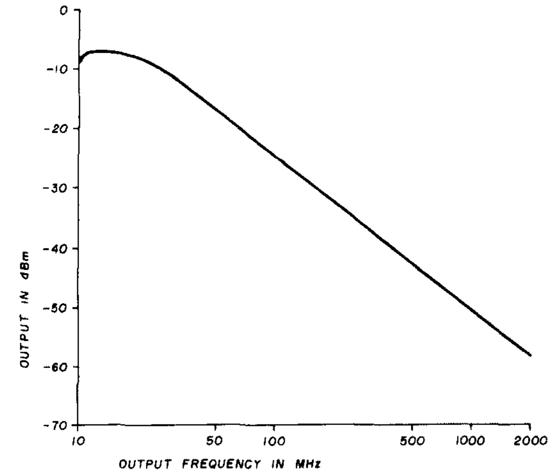


fig. 7. Typical output power versus frequency from a secondary frequency standard comb generator.

construction

The entire circuit including battery can be built *on the cover* of a Pomona model 2901 or equivalent cast box. First a double-clad printed circuit board about 1-7/8 × 3 inches (48 × 76 mm) is attached to the cover of the box and held in place by the output connectors, switch, trimming capacitor, and LED. A suitable battery clip is mounted on the remaining available space on the box cover. Two insulated standoff terminals are used to hold the crystal leads and associated temperature compensation capacitors. The wiring is done point-to-point in space, as suggested in previous articles.[1]

initial testing and calibration

If you don't want to go through the temperature calibration procedure described, just use 10 pF N750 capacitors for C2 and C3 and you should be very close. A small 2-meter quarter-wave pull-up type whip makes a suitable radiating antenna for the entire spectrum of interest. The calibrator is now ready for testing. This can be easily done by listening for the 7th harmonic on a 28-MHz HF receiver (the signal may not be strong at this frequency) or the 36th harmonic at 144 MHz on a 2-meter receiver. First peak C6 for maximum output on a 2-meter or higher receiver. Next, vary the calibration trimmer, C1, and check that the frequency varies a few hundred Hertz and that it is approximately correct. If the frequency is too low, either C2 or C3 will have to be decreased by perhaps 5 pF.

To verify adequate temperature compensation, you'll have to measure the output, preferably with an accurate frequency counter, or zero-beat it with WWV at 20 MHz using a suitable HF receiver. The entire circuit can then be placed in a refrigerator, (typically about 41 degrees Fahrenheit or 5 degrees Centigrade). Measure the frequency after it's been in the refrigerator for 15 to 30 minutes. Compare the reading to the one made earlier at room temperature. After a few tries, you will see if the calibration is adequate. If the temperature drift when the calibrator is cold causes the frequency to go *below* the room reference frequency, too much compensation is being used, and vice versa. (See **fig. 5** for further information.) Also, whenever you change a capacitor, allow at least 15 minutes for its temperature to stabilize before taking data.

Zero-beating to WWV can be tricky. It's best to move the calibrator around the room until the proper amount of injection is experienced. Also try to zero-beat during the interval when the tones are not present (usually the last 15 seconds of each minute). Watching the "S" meter can be quite helpful. When close to zero-beat, the meter will waiver noticeably and the beat will sound like a chirping canary. When the meter moves once per second, *you are within 1 Hertz* of the correct frequency. One beat every 15 to 30 seconds is very satisfactory. Be aware that because of atmospheric effects the accuracy of WWV on the HF spectrum is good only to about 1 part per hundred million. However, this should be

short circuit

VHF/UHF World

A chart in fig. 9 (page 88) of W1JR's "VHF/UHF World" column in April, 1984, made reference to "Note 1." Unfortunately, Note 1 was not included at the bottom of the figure, as the author intended. Note 1 should read, "Not used on this frequency."

more than sufficient for Amateur work.

using the calibrator

At first, it will be worthwhile to test the accuracy of the calibrator several times during the first few months of operation because some aging will inevitably occur. This will also let you develop confidence in its use. After initial burn-in, I usually test calibration against WWV about once a year. One caution is advised: *Do not place the calibrator where it will be subjected to temperature variations such as on top of another piece of electronic gear.* I usually place mine on a table or desk one or two meters away from the receiver in use. This yields more than adequate injection even at 2304 MHz. Typical warmup is 15 seconds. To conserve battery life, don't forget to turn the calibrator off when not in use!

summary

Even though the state-of-the-art in accurate frequency determination has advanced significantly in recent years, a secondary frequency standard is still useful for verification of true frequency. The calibrator just described is easy to construct and adjust, relatively inexpensive, very stable, and portable. I now have six of these units operating on different frequencies and always carry one along on expeditions. Once you have one, you'll never want to be without it.

references

1. Joe Reisert, W1JR, "VHF/UHF World: VHF/UHF Receivers," *ham radio*, March, 1984, page 46.
2. Mike Metcalf, W7UDM, "A VHF-UHF Marker Generator," *VHFer Magazine*, May, 1965, page 3.
3. Irvin M. Hoff, W6FFC, "The Mainline FS-1 Secondary Frequency Standard," *QST*, November, 1968, page 34.

VHF/UHF coming events

October 6: *Mid-Atlantic States VHF Conference, Warrington, Pennsylvania. (Contact W3CCX for information.)*
October 6-7: *International Region I UHF/SHF Contest. (Contact RSGB for information.)*
October 20 and 21: *ARRL International EME Contest* (second weekend).
October 20: *Predicted peak of Orionids Meteor Shower at 0515 UTC*
October 24: *EME Perigee*

ham radio

a high-efficiency
top-loaded vertical

You don't need traps, coils, or vast acreage to get good performance on 160, 80, or 40 meters

The efficient, easy-to-build antenna described in this article requires no radials, traps, or loading coils. The reader with an appetite for low-band DX should be able to duplicate or better my results on 160 meters and, by appropriate scaling, on 80 and 40 meters as well.

Like any serious low-band DXer, I wanted to join the "big boys" using a vertically polarized signal because it has long been demonstrated that horizontals, unless raised to astronomical heights, are unable to do much more than take up space at the bottom of the pile-ups.[1] The customary current-fed, 1/8- or 1/4-wave type verticals require an immense radial ground system to be efficient and competitive — yet my terrain consists of deep ravines, gullies, and cliffs that make the installation of radials, or even elevated counterpoise systems, virtually impossible. To add to the misery, the soil at my QTH is almost 100 percent hard sandstone and as such approaches pure silicon in conductivity. Obviously a "non-standard" approach to an antenna was necessary. After reviewing many texts and articles I decided that some form of "ground independent" antenna would be required, particularly one that had its feed point at ground level. The "Bobtail Curtain"[2,3] version of ground-independent systems met the basic requirements, but for 160 or even 80 meters the dimensions become truly heroic; in addition, my requirement was for unidirectional transmission. My first step was to contact my old friend Woody Smith, W6BCX, who had fathered the fabulous Bobtail, in order to get firsthand information on the "care and feeding" of this type of system.

My call to Woody obviously struck a responsive chord and after a few thousand well-chosen words and equations, we agreed upon the novel form of vertical that this article describes. Sparing you the background — which covered inverted ground planes, slopers, center-fed loaded dipoles, and of course the Bobtail itself — the general parameters decided upon were that the new vertical design should require supports no higher than 70 feet (21.34 meters) for the 160 meter version or 35 feet (10.67 meters) on 80; should not need more than 95 feet (28.96 meters) spacing between supports on 160 and half that on 80; should be highly efficient in power transfer; should exhibit a high impedance feed point at the bottom of the vertical radiator and should be relatively easy to construct

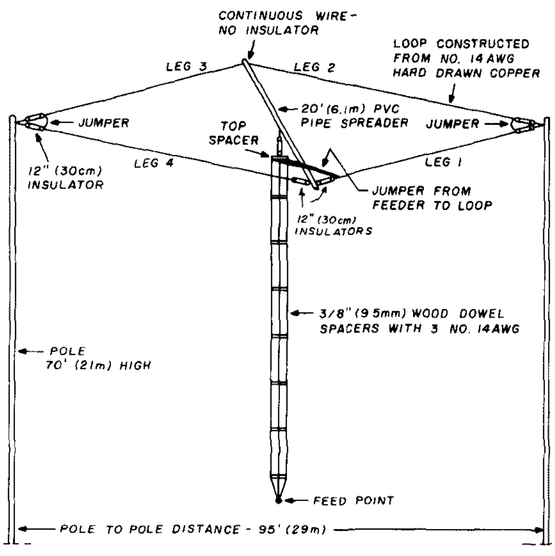


fig. 1. Efficient top-loaded vertical dimensioned for 160-meter operation.

By Howard F. Shepherd, W6US, c/o Hypercom, P.O. Box 768, Del Mar, California 92014

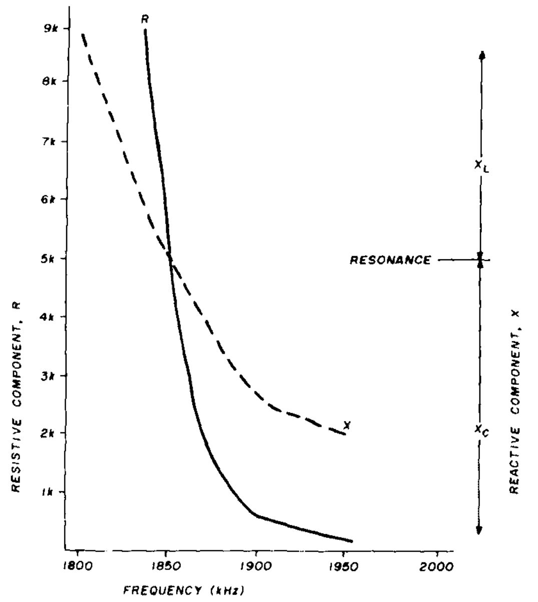


fig. 2. Plot of resistive and reactive component of 160-meter top-loaded vertical versus frequency.

item	description
L1	40 turns No. 10 AWG spaced 0.2 inch per turn, 4 inch diameter
L2	7 turns No. 10 AWG spaced 0.2 inch per turn, 5-1/4 inch diameter
C1	two 500 pF, 2000 volt variable capacitors in parallel
C2	500 pF, 500 volt variable capacitor with 0.002 μF mica in parallel
C3	100 pF, 20 kilovolt fixed vacuum capacitor

Note: For upper portion of 160 meter band, L1 is tapped down 8 turns by HV relay and 0.002 μF mica is disconnected from C2 by another relay.

and test using easily obtained materials. How did it work out? If you're not one of the fortunate few who already have a full 1/2-wave vertical on 160 complete with at least 120 radials or a solid silver-plated copper ground sheet of comparable size, read on!

Fig. 1 shows the physical layout of the W6US 160-meter vertical. No. 14 AWG copper wire is used throughout. The vertical radiator uses three wires spaced 1 foot (30.48 centimeters) apart by 3/8-inch (0.95 cm) diameter wooden dowels. Because the wires are at the same potential, no special treatment is required for the dowels electrically, but a coat of varnish or wax enhances their weatherability. The use of a four- or six- wire cage feet (1.22 meters) in diameter would be helpful as far as increased bandwidth is concerned; however, unless the effective conductor diameter of the diamond-shaped single-turn top loop is also increased by paralleling wires, don't expect to greatly widen the frequency response.[4] It should be noted that if the vertical radiator is so increased in electrical length, the physical length of the loop will have to be decreased, and provision made for the increased weight of the cage system.

how does it work?

The general theory of operation presents little mystery, but despite the W6US system's resemblance to the usual "top-loaded" short vertical, the reader will note that in essence the feedpoint is at the end of an electrical 1/2-wave of conductor and represents a point of high radiation resistance and impedance. This becomes paramount when the efficiency of this antenna is compared to that of the usual top- or bottom-loaded systems.[5]

$$\text{antenna efficiency} = \frac{R_{rad} \times 100}{R_{rad} + \text{all circuit resistance}} \qquad (1)$$

where R_{rad} = the radiation resistance of the antenna and "all circuit resistance" represents the resistance of the ground system, the loading inductors, wire resistance and *any other resistances present which dissipate power as heat.*

A resonant 1/4-wave vertical exhibits a R_{rad} of about 36 ohms with 10 ohms representing the total resistance of the ground system, coupling circuit, and other loss elements. Inserting these two numbers in **eq. 1** produces an efficiency of 78.26 percent. On 160 that 1/4-wave vertical would be about 130 feet (36.92 meters) tall. If instead a 60-foot vertical (1/8-wavelength high) is used, the R_{rad} drops to around 4 ohms and the efficiency diminishes to 28.57 percent![6] For 100 watts transmitter output, only 28.57 watts would be radiated — about one-third of the radiation from the full size 1/4-wave vertical, assuming the losses were identical (which is really not the case because the shortened vertical would require more coupling inductance and consequently more power would be consumed in the coupling circuit instead of reaching the antenna).

This is where the W6US ground independent system enters the race with a big lead. From **fig. 2**, which depicts the variation in measured antenna resistance and reactance of the system shown in **fig. 1** note that at resonance — i.e., zero reactance — the resistance is close to 5000 ohms. Applying this number in **eq.1** and assuming for the moment that the "all circuit resistance" figure stays at 10 ohms, the efficiency is 99.8 percent — almost four times the indicated

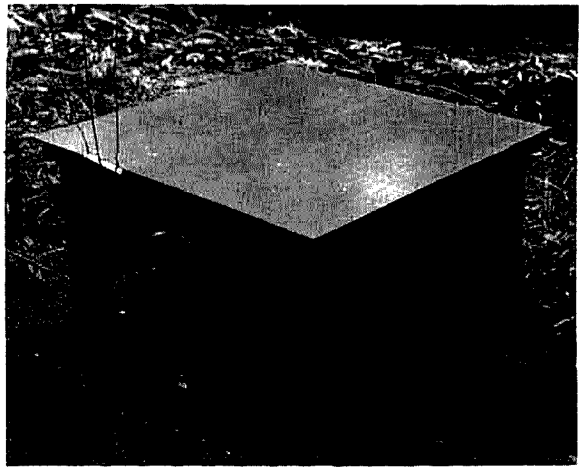


fig. 3. Coupler houses high voltage matching circuit components.

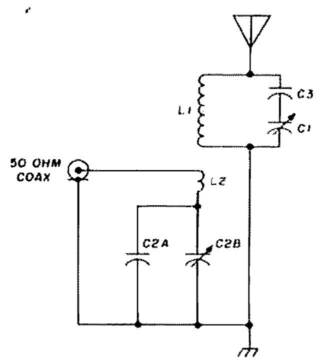


fig. 4. Coupler provides match between high impedance seen at antenna terminals and 50-ohm coax line.

efficiency of the usual 60-foot vertical radiator! This type of effect is exactly what makes the Bobtail and its other cousins such real winners. Best of all, this is achieved without an extensive ground radial system. In this case a 10 × 10 foot (3.05 × 3.05 meter) square of galvanized poultry wire is all that is necessary. A 1/2-inch (1.3 cm) 8-foot (2.43 meter) copper pipe is driven into the ground at the center of the wire screen for lightning protection. The small "coupler house" is placed on this screen as shown in **fig. 3** with the ground portions of the coupling circuit strapped to the pipe, which in turn is bonded to the poultry wire.

The resistance/reactance graph shown in **fig. 2** illustrates that these curves are steep with respect to frequency. Obviously, this antenna is very high "*Q*" and the bandwidth of antenna plus coupler is 6.7 kHz on 160 between the 1.2:1 VSWR points. Because of the narrow bandwidth, I provided for remote adjustment of the tuning capacitor in the coupling network with a 90-volt 60-Hz selsyn pair. The complete coupler circuitry, a parallel tank circuit consisting of L1 and C1 in series with C3 which resonates at the desired frequency, is shown in **fig. 4**. The tank is inductively coupled to the series circuit L2 and C2 into which the 50-ohm coax feedline terminates.[7] Not shown in **fig. 4** is a relay switching arrangement whereby L1 is tapped down and the value of C2 is decreased to provide for a wider total tuning range on 160. The interior view of the coupler housing in **fig. 5** shows the location of all the components. Remote adjustment of the coupler is obtained by having a selsyn drive the capacitor C1 via a reduction ratio chain drive which allows the use of a ten turn dial on the selsyn in the shack. The dial includes a small chart that provides a quick dial versus frequency reference for setting.

The use of the C1-C3 series combination in the

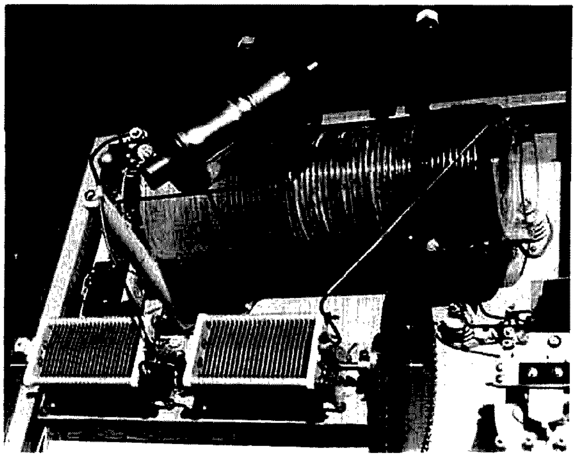


fig. 5. The interior view of the coupler housing illustrates the precautions taken with component placement as a result of high voltages present.

coupler was dictated by the very high voltage developed at the base of the antenna at full legal power input. C3 is a 100 pF, 20,000-volt fixed vacuum capacitor that receives most of the voltage across the tank because C1 is varied from its full capacity of1000 pF to about 300pF to cover the tuning range. This combination also provides some degree of bandspread, helpful in view of the sharpness of tuning. I would suggest that you use high quality insulation, and plenty of it, both at the coupler and diamond loop ends of the antenna. With the impedance ranges involved, the voltages are awesome at all but the lowest power input.

adjusting loop to resonance

Once you've assembled the antenna and set it in place, the next step is the adjustment of the length of the diamond loop to achieve resonance. It is assum-

ed that you will have set the overall length to a half-wave according to the usual equation; $468/f_{MHz}$. For 1850 kHz this figures out to 252.97 feet (77.11 meters). There are a variety of ways to determine whether resonance has been achieved, ranging from loosely coupling a grid dip oscillator to the bottom of the vertical radiator to using an accurate RF impedance bridge, which was the method I used (see **fig. 6**). Another method that is quite easy to use is to resonate the tank circuit in the coupler using a GDO to your desired frequency without having it connected to the antenna. Then attach the antenna temporarily about halfway up the tank inductor from ground. If the antenna is resonant at that frequency, the tank circuit will still resonate with the GDO at the same setting. If not, as almost always is the case, you will have to adjust the GDO to find the dip again. Then try a different diamond loop length and go through the process once more. By noting the movement of the dip, you can easily tell if the antenna length adjustment was made in the right direction. By successive tries you'll find a length for the diamond loop that will closely approximate resonance. The antenna is then connected to the hot end of the tank circuit for adjustment of the VSWR and operation.

It is well to note that the above antenna pruning method can also be used to determine the harmonic resonance by setting up a temporary parallel tuned circuit that will resonate on the harmonic. If this antenna is to be used on both 160 and 80, or 80 and 40, you may wish to choose resonant frequencies that cover the band segments of interest. As shown in **fig. 2**, my antenna has its primary resonance at about 1854 kHz. Testing has shown its second harmonic to be close to 3949 kHz.

fig. 6. W6US determines the verticals resistive and reactive components with a General Radio GR1606A RF bridge.

If you are seriously interested in this type of ground-independent antenna, I would suggest some reading in books and articles that deal with VLF antenna systems.[8] Some of the earlier systems closely parallel the dimensions of the antenna described here. I would also be interested to learn of your experiences with the W6US antenna on other bands — for example, on 10 meters, where the vertical radiator would be a magnificent 4.32 feet (1.32 meters) high and the total span about 11.88 feet (3.62 meters)! Also left to your experimentation is the question of whether the addition of a conventional radial system might enhance this antenna for DX.

The sticky point in this investigation is probably whether the raising of the current maximum by the ground independent approach lessens the need for radials to lower the angle of radiation. The Bobtail does not seem to profit by them, but here the vertical radiator is only half as long. At least it seems well established that the increased efficiency goes a long way toward successful long-haul DX.

testimonial

No antenna article is complete without a relatively rosy exposition of the superlative reports received and the choice DX worked. To maintain tradition, I will therefore note that compared with a great variety of 160-meter systems previously used, this W6US ground independent system has been a real winner for me. A substantial number of solid QSOs have been logged with stations 500 miles or more distant while running just 10 watts transmitter output on SSB. Believing that more of a good thing is even better, I am presently working on a second version of this antenna. I sincerely hope that those who try this approach will provide still more data to benefit us all.

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ham radio

extended/expanded power dividers

Try copper water pipe, hobby-shop brass tubing for quality RF components

In weak signal VHF/UHF work (EME and various scatter modes)* nothing can be more frustrating than the inability to read faint signals you know are present but are largely obscured by noise.† The problem is often solved by increasing the size and/or number of antennas; when properly combined in a low loss system, doubling the number of antennas will yield very close to the theoretical 3 dB gain. But there is little to be gained in capturing a weak signal unless it is nurtured with great care on its way to the preamplifier. To do this, one should use the best quality feedline, cut as short as possible, to the power dividers. Ideally, the power dividers should be constructed to facilitate the use of short lines; unfortunately, most are either 1/4 or 1/2 wavelength long, as illustrated in **fig. 1**. With stacking distances between antennas as great as two wavelengths, these dividers do not permit the use of short feedlines.

theory

While most readers are probably familiar with the power divider basics, this section has been included for the benefit of those who may not be, and for the further purpose of showing the evolution of extended power dividers.

Power dividers use the *"Q"* section used as a means of matching the load of more than one antenna to the feedline. A quarter wavelength of feedline will match the impedance of feedlines of the same type (coaxial or parallel) but having different impedances when the impedance of the matching section is: $Z = \sqrt{Z_1 Z_2}$ and is known as a *"Q"* section. (This discussion is limited to coaxial feedlines; all loads are assumed to be equal.)

When two or more loads, in the form of antennas or other power dividers, are fed from one transmission line, they appear to be in parallel. Four 50-ohm loads represent a combined load of 12.5 ohms. In order to arrive at the impedance of the matching section we divide the impedance of one load by the number of loads, and this becomes Z1. Z2 will be the impedance of the input feedline. If the matching section is to be divided into two or more branches, as shown in **fig. 1B**, then its impedance should be multiplied by the number of branches so that the combined value is the originally calculated impedance.

design

The power divider shown in **fig. 2** is a combination of the two types of dividers shown in **fig. 1**. The center portion of **fig. 2**, section A, is a 1/2-wave power divider or matching section that matches the input impedance to the impedance of the two B sections. In theory, section B could be any impedance that would fit into the overall design, but as a practical matter it should be close in value to the output impedance. The two C sections at each end are 1/4 wave dividers that match section B to the two output ports at each end. I am presently using two of these in my 144 MHz EME array and have been more than satisfied with their performance.

A different method of accomplishing the same thing is shown in the power divider illustrated in **fig. 3**. Simpler in design, and easier to construct, this one

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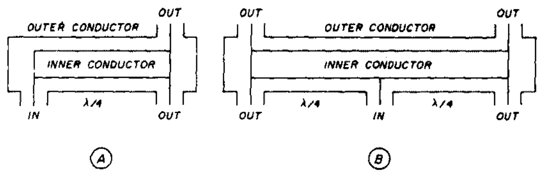
*See "The VHF/UHF Primer: An Introduction to Propagation," by Joe Reisert, W1JR, *ham radio*, July, 1984, page 14, for a comprehensive survey of the VHF/UHF modes.

†HFers experience that same frustration. — **Editor.**



Note 1: 1/4 wave power divider:

$Z = \sqrt{Z_{IN}Z_{OUT}/N}$

where Z_{OUT} = impedance of one output port,
and N = the number of output ports.
Center conductor = λ/4 = $2952/F_{MHz\text{-}IN}$

Note 2: 1/2 wave power divider:

$Z = 2\sqrt{Z_{IN}Z_{OUT}/N}$

where Z_{OUT} = impedance of one output port,
and N = total number of output ports.
Center conductor = λ/2 = $5904/F_{MHz\text{-}IN}$

Note 3: Impedance of air dielectric coax with round outer and inner conductor:

$Z = 138 \log \frac{D}{d}$

where D = I.D. of outer conductor
and d = O.D. of inner conductor

Note 4: Square outer, round inner:

$Z = 59.96 \ln (1.0787 \frac{W}{D})$

(See ham radio, March, 1981, page 64, Meyer, W6GQV)
where W = inside dimension of one side of square outer
and D = O.D. of inner conductor.

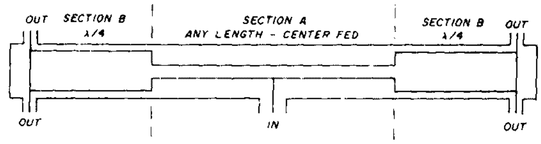
fig. 1. Forms of conventional power dividers. (A) 1/4 wave power divider, and (B) 1/2 wave power divider.

should perform equally as well as the others. At each end of a 1/2 wave four-way power divider, the impedance is one-half the impedance of each output port. This will be 25 ohms for 50 ohm feedline or 37.5 ohms for 75 ohms feedline. We simply extend the matching section, at each end equally, to the desired length, at the impedance of $Z_{OUT}/2$. (I will be using this design in my 432 MHz EME array now under construction.)

common materials are all that's needed

I use copper water pipe, straight and hard drawn, for the outer conductor of my power dividers. It is available in three wall thicknesses for each nominal diameter: type K, thick wall: L, medium wall; and M, thinwall. **Table 1** shows the nominal size, O.D., and I.D. of copper pipe in sizes from 1/2 inch (12.7 mm) through 1-1/2 inch (38.1 mm). For the center conductor, I use brass tubing, available in hobby shops, in diameter increments of 1/32 inch (0.8 mm) from 1/16 inch (1.6 mm) through 21/32 inch (17 mm), having a wall thickness of 0.014 inch (0.36 mm) and sold in lengths of 1 foot (30 cm) or 3 feet (91 cm). It is easily spliced by soldering a short section of the next smaller diameter as an internal sleeve.

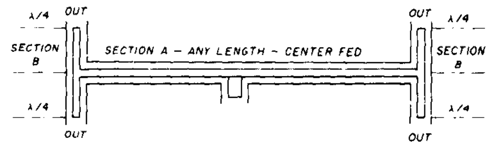
Square aluminum tubing is also good for the outer conductor, but I haven't been able to find it in lengths greater than 8 feet (2.4 meters), which isn't long enough for widely spaced 2-meter arrays, though it is easy to work with. Coaxial cable connectors can be held in place with small sheet metal screws, or the tubing may be drilled and tapped.



Section A: Z = 100 ohms D/d = 5.3 Section B: Z = 50 ohms D/d = 2.3

Note: The 50-ohm input is split into two 100-ohm loads. The "Q" section (Section B) transforms the 100-ohm portion to the 25-ohm load presented by the two 50-ohm outputs at each end. This design is NOT suitable for 75-ohm feed systems, for reasons explained in the text.

fig. A1. 50-ohm 4-way power splitter.



Note. The impedance of this power divider is 100 ohms throughout and provides a perfect match for 50 ohms in, 4 × 50 ohms out. While the same design for 75 ohms appears satisfactory on paper, it is NOT so in the actual use.

fig. A2. The "One Hundred OHM" 4-way 50-ohm power divider.

good performance requires careful construction

In layout, measurement and assembly, accuracy and uniformity are essential. After first determining the desired overall length, find and mark the exact center of the outer conductor, making a very light cut around the circumference with a tubing cutter. At points equidistant from the center, and 3/4 inch (19 mm) from each end, make a similar mark. These are the points at which the connectors or sleeves will be located, as well as the center and each end of the center conductor. Drill holes 180 degrees apart, on opposite sides of each end of the outer conductor, and one at the center, to accept the coaxial cable, connectors or sleeves. The center hole should be either in line with two end holes on one side, or at 90 degrees, depending on the orientation of your feedline. A 3/8 inch (9.5 mm) hole is drilled to one side of the center hole, just close enough that it will not run out and overlap the center hole. This is for the purpose of being able to have access for soldering the center conductor to the coaxial connector.

In preparing the different sections of the center conductor, the smaller diameter piece should be long enough for about a 3 inch (76.2mm) overlap into the larger size tubing to which it is to be joined. A reducing sleeve is made by telescoping together short sections of brass tubing, about 3 inches (76.2mm) long, so as to fill the space between the smaller and larger pieces

to be joined, and to keep them straight and concentric. If the center conductor is to be made in one piece, insulating spacer washers should be put in place before the sections are soldered together.

I use teflon spacer washers* to prevent sag in the center conductor and to maintain concentricity. These are placed on the smaller diameter portion at each point of change in diameter, at the center and on each end. The spacers should fit tightly on the center conductor and loosely in the outer conductor. My center conductors were assembled in two pieces divided at the center. This permits setting the spacers in place on the smaller diameter portions after the soldering is done. They are joined in the center with an internal sleeve after being inserted in the outer conductor. The center conductor should be cleaned and polished after assembly and handled with paper towels or tissue during insertion in the outer conductor. If the inside of the outer conductor is not bright and shiny it should likewise be polished. A small notch or "Vee" should be cut out of each teflon spacer, at the perimeter, to permit the flow of air through the power divider; this is important to prevent the accumulation of moisture.

After inserting the center conductor into the outer conductor, the coaxial connectors are ready to be installed. I recommend the use of UG-58/U connectors with teflon insulators. (Other types of insulation will not withstand the heat generated during the soldering process.) The flange of the UG-58/U should be removed — with a lathe, grinder, or otherwise — to facilitate soldering. A 7/16 inch hole in the outer conductor will provide a neat fit for the portion of the UG-58/U below the flange.

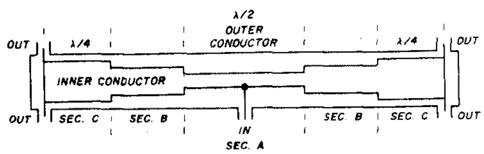
I use coaxial connectors only at the center or input port. At the output ports, I use copper sleeve couplers sized to fit 3/8-inch (9.5 mm) copper pipe. They are about 0.502 inch (12.75 mm) I.D. and provide a very good fit for the shield of the RG-331 hardline leads to the antennas. The sleeves are tinned on the inside. A small stainless steel wormscrew hose clamp is put around the sleeve after the hardline is installed.

After all soldering, assembly, and installation have

table 1. Nominal size, O.D. and I.D. of different grades of copper water pipe.

nominal size (inches)	O.D. (inches)	I.D. type K (inches)	L (inches)	M (inches)
1/2	0.625	0.527	0.545	0.569
5/8	0.750	0.652	0.666	0.690
3/4	0.875	0.745	0.785	0.811
1	1.125	0.995	1.025	1.055
1-1/4	1.375	1.245	1.265	1.291
1-1/2	1.625	1.481	1.505	1.527

*Available from the author at nominal cost.

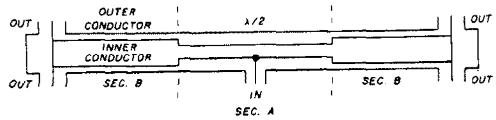


Note: This power divider may be any length greater than one wavelength. Section A is a two-way 1/2 wave power divider. $Z = 2\sqrt{Z_{IN}Z_B}$. Sections B may be any length, but must be equal. The impedance may be any value within practical limitations. Sections C are 1/4 wave two-way power dividers. $Z = \sqrt{Z_B Z_{OUT}/2}$ With an outer conductor of 1-inch type L straight, hard drawn copper water tube (1.025 inch I.D.) impedance of sections A, B, and C and diameter of center conductors will be:

Performance parameters of multisection power dividers.

Z_{IN} (ohms)	Z_{OUT} (ohms)	section A Z (ohms)	section A diameter (inch)	section B Z (ohms)	section B diameter (inch)	section C Z ohms	section C diameter (inch)
50	4 × 50	71	5/16	51	7/16	36	9/16
50	4 × 75	85	1/4	71	5/16	51	7/16
75	4 × 75	102	3/16	71	5/16	51	7/16

fig. 2. Long power divider.



Note: Section A has the same impedance as a four-way power divider, as illustrated in fig. 1B. In the case of 4-way dividers of this type the simplified formula is $Z = \sqrt{Z_{IN}Z_{OUT}}$, where Z_{OUT} is the impedance of one output port. Sections B may be any length and of equal dimension. The impedance is $Z = Z_{OUT}/2$. Typical applications:

Z_{IN} (ohms)	Z_{OUT} (ohms)	section impedances (A)	(B)
50	4 × 50	50	25.0
50	4 × 75	61	37.5
75	4 × 75	75	37.5

fig. 3. Another form of long power divider.

been done, the ends and the center access hole should be sealed. It is not necessary to solder the coverings in place, but they should be held in place with a good sealant and wrapped tightly with stretch tape. All points of possible moisture entry — even the solder joints — should likewise be sealed.

The dimensions given in **fig. 2** will yield impedances very close to the calculated ideal. Preliminary calculations indicate that 3/4 inch (19 mm) type M tubing would be a better choice for outer conductor in the **fig. 3** design, using brass tubing for the center conductor.

what to expect

It is said that 1 dB is the least amount of change in sound that can be detected by the human ear. In a noise-free environment, this is probably correct. But in EME work, I question its accuracy. Often we can detect a change in the pitch of the "white noise" as we tune our receivers across a given frequency. Even though there's a signal present, the dits cannot be distinguished from the dahs. Perhaps if this were 0.9 dB, the signal would jump right out of the noise and be readable if we could supply the additional 0.1 dB? The low-loss feed system described herein may make such a thing possible.

appendix

other designs for power dividers

In each of the power dividers previously described, a "*Q*" section is employed at the input port. This is not necessary in all cases. The input may be split into two branches, each of which is twice the impedance of the input feedline, with no resulting mismatch. The "*Q*" sections may be placed at each end, in either 1/4 or 1/2 wave form.

Figures A1 and **A2** illustrate this design in each of these forms. I have designated the first device as a "splitter" to distinguish it from the one in **fig.** 3. This design is not suitable for a 75-ohm feed system, which would require splitting into 150-ohm branches. The D/d ratio for 150 ohms would be 12.2, resulting in a center conductor much too small to be practical.

The "hundred OHM" (**fig. A2**) divider is particularly appealing to me, because it involves no change in diameter of the center conductor. The idea for this one occurred to me about a month ago, and I hastily assembled two of them for my 2-meter EME antenna (8 x 2.2 wavelength NBS Yagis). Initial results are very encouraging: SWR is very good, and my own SSB echoes from the moon were readable with a sky temperature of about 400 degrees K. (Below 300 degrees K is considered "quiet sky.")

construction details for the "hundred OHM" divider

Outer conductor: 3/4 inch (19 mm) copper pipe, type D, 0.833 *inch (21.2 mm) I.D., if available. Type M, 0.811 inch (21.2 mm) I.D.,* is satisfactory substitute.

Inner conductor: 5/32 inch (4 mm) brazing rod or brass tubing. Tubing is preferred because of its usefulness as internal coupling sleeves.

Center and end Tee's: 3/4 inch (19 mm) copper plumbing Tee's.

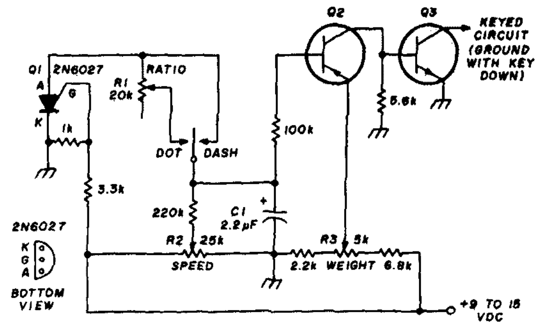
Center conductor splices: 5/32 inch (4 mm) I.D. brass tube. Miniature Tee's fabricated from same material at each end. (I used brazing rod center conductor.)

Coax connectors: UG-58 with teflon insulation, flanges turned to fit pipe, soldered.

ham radio

the weekender

simple, compact QRP keyer



Q2 PNR SILICON TRANSISTOR 2N3906 OR SIMILAR

Q3 NPN SILICON TRANSISTOR 2N3904 OR SIMILAR

Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

fig. 1. Simple QRP keyer uses unijunction transistor for timing.

I recently built a **QRP CW transceiver** and decided to use one small cabinet to house the radio, rechargeable batteries, and a keyer. After completing the transceiver and squeezing in the batteries I found I had left myself less than two square inches of board space for the keyer circuitry.

Because contemporary electronic keyers use one or more ICs and quite a few additional components, I needed to find a simpler, more compact approach. I also wanted self-completing characters and very low battery drain. Dot and/or word memory and digital precision were features I could do without in this application.

The circuit that evolved, shown in **fig. 1**, does the job very nicely, fits easily on 1.5 square inches of board space, and consumes less than 4 mA, key-down, at 12 VDC.

In lieu of the usual digital clock, this keyer uses one of the unijunction family of devices as the basic timing device. Unijunction devices (UJTs), which have been around almost as long as the basic transistor, have several unique characteristics that make them quite different from those of the conventional two-junction transistor. These features include stable triggering voltage, very low value of firing current, high pulse current capability, and low cost.

The circuit (**fig. 1**) uses these characteristics of the UJT to generate a precisely timed sawtooth waveform. This is formed by the exponential voltage build-up across capacitor C1, and the abrupt discharge of this voltage when the UJT "fires," or triggers. Dashes and dots are self-completing because the UJT has no effect on the capacitor charge until the triggering point is reached. Potentiometer R2 controls the rate of charge of C1 — hence the keying speed. R1 sets the ratio for the generation of dots.

The sawtooth waveform is applied to a PNP transistor, Q2, where it is shaped and amplified. The output of Q2 is direct-coupled to Q3, the switching or keying transistor. R3, the "weight" potentiometer in the emitter of Q2, controls the switching threshold (the on-off periods) of Q2 and Q3.

I use the collector of Q3 (which goes to ground when the key is down) to key the emitter of the transmitter driver transistor that draws about 20 mA. You can, of course, key any circuit that requires ground to transmit provided you stay within the current and voltage capabilities of Q3.

A relay with a DC coil resistance of 500 to 3000 ohms can be used in the collector of Q3. The relay coil is connected between the collector of Q3 and +12 VDC as shown in **fig. 2.** This arrangement will increase the current drawn by the keyer another 5 to 25 mA, depending on the relay coil DC resistance. Reed relays are recommended and work well for keying small, non-inductive current loads. An inductive load such as an iron-core choke in series with the keying lead will generate an inductive charge. If not damped, this charge will quickly weld together the contacts of the reed relay.

limitations

The limitations of simple circuits such as this are often overlooked or glossed over by enthusiastic authors, with resultant heavy mail and telephone traffic to the publication and/or author. To minimize such communication, let me say that this simple keyer will work very well if the following precautions are taken.

First and most importantly, the DC supply voltage must be constant (either through regulation or a stable source) because the timing of the sawtooth generator as well as the switching thresholds of Q2 and Q3 are a direct function of the supply voltage. Poor voltage

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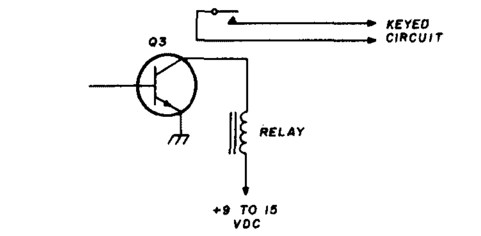

fig. 2. Reed relay can be driven by Q3 for keying circuit isolation.

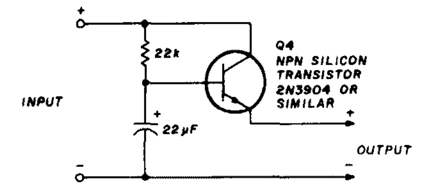

fig. 3. Decoupler circuit helps to reduce AC power supply ripple from DC line.

regulation will cause erratic characters. Long-term changes in supply voltage from, for example, 12 volts to 10 volts, will drastically affect the weight of the characters, and to a lesser extent, the speed. The keyer will work well with supply voltages of 9 to 15 volts. Once the operating voltage is selected it must remain at that value; if it does not, all controls will require re-adjustment.

I use heavy duty batteries to avoid these problems. If an AC supply is used, the simple regulator-decoupler shown in **fig. 3** does a good job of cleaning up the supply voltage source.

With the possible exception of C1 and Q2, component tolerances are not critical. C1 must be 2.2 microfarads. (1µF is too small and 3.3 µF is too large.) Q2 should have a fairly high beta — at least 60 or 70 — so either avoid the unknown junk-box types entirely or use a socket and make sure your choice works properly. The 2N6027s appear to be non-critical. All six of mine, from two different manufacturers, worked well with no noticeable variation in the characteristics needed for this application.

Finally, if a relay is used, try to use a reed type or one with a short, snappy armature throw. The keying waveform coming out of Q3 is more triangular than rectangular. A sloppy relay will generate satisfactory dashes but skimpy dots, and you won't be able to compensate for this with the "weight" or "ratio" controls.

ham radio

applied Yagi antenna design part 6: the model and a special teaching tool

Understanding and using the FORTRAN program

In this concluding article, the FORTRAN code used in the model is discussed on terms of its organization and logic flow, and an interesting teaching example also presented.

A copy of the program is available from *ham radio* (send an SASE with 37¢ postage). Please note that the program, designed to run on a mainframe computer, requires a FORTRAN 77 compiler or its equivalent. While a knowledgeable programmer can convert this program to BASIC, an understanding of the "architectural" differences between FORTRAN and BASIC is essential, as is familiarity with the use of both radians and degrees, as well as the rules of complex (vector quantity) arithmetic. A careful rereading of the first two articles in Lawson's series[1,2] is also recommended.

FORTRAN program structure

The program consists of seven logical parts, as depicted in **fig. 1**. The Driver contains array definitions, constant declarations, and the control logic for the five main subroutines. Each subroutine's name is an accurate description of what it does. The mathematics library contains the functions used in computing impedance components. The model's code has been carefully written to minimize execution time. This is especially true for the reactance calculations because they can be executed tens of thousands of times in a single session of Yagi iterations.

*We regret that neither the author nor the staff of *ham radio* can provide software assistance or consultation.

The Driver contains the Poynting vectors necessary for pattern calculations. These vectors are listed as constants in an extended DATA statement. As the vector values for the second 90 degrees follow from those of the first 90 degrees, a DO loop makes this assignment. This same loop adjusts the vector values slightly so a dipole has a calculated gain of 2.14 dBi. Further manipulation of the 0.997 multiplier leads to degraded pattern calculations. This is followed by calculation of the dipole pattern constants for each of 180 degrees. All of the Driver code to this point is executed only once — at the start of the initial run. Now the five subroutines can be called to calculate Yagi performance parameters. If a series of iterations similar to those in the first five parts of this series is desired, the nested DO loops begin at this point.

For those articles a primary loop was used to iterate reflector length. A secondary loop iterated director length within each discrete reflector measurement. For tapered directors a tertiary loop was used to apply the tapering constant to the second and subsequent directors. By definition, the first director cannot be tapered. If spacing is to be iterated, the loops for spacing can also be placed here, either in place of the length loops, or within them. When a frequency scan of a specific Yagi design is desired, element length, spacing and diameter are set as constants, and a frequency DO loop is coded. Element diameter (ELDIAM) is usually a constant, but this parameter can also be iterated.

GETDATA originally served to prompt for and accept the individual Yagi design approach parameters. After obtaining calculations for the Yagis in use at WB3BGU, the decision was made to write this six-part

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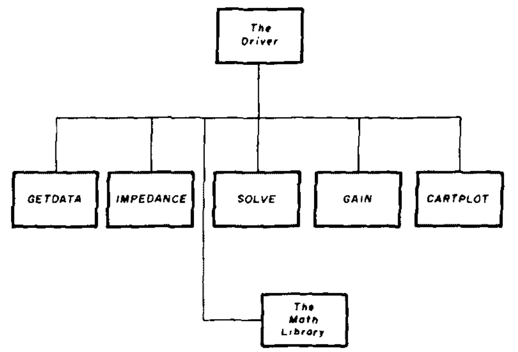

fig. 1. Internal organization of the Yagi analysis program.

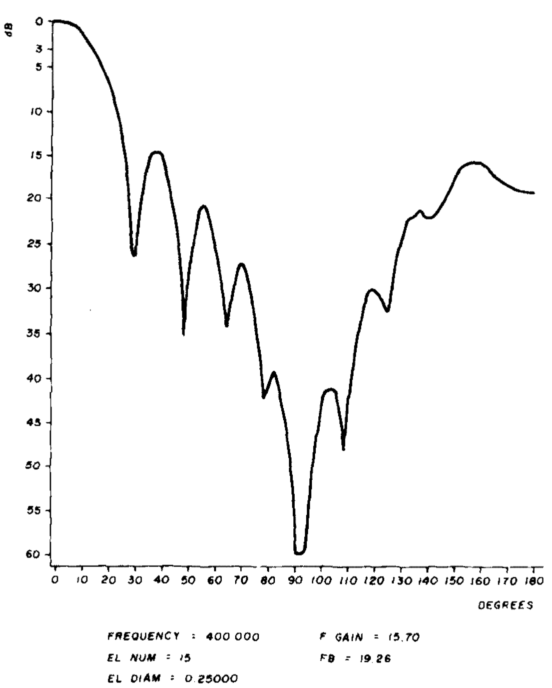

fig. 2. E-plane plot of the NBS 4.2 wavelength Yagi.

series of articles. Hence the need arose for either massive amounts of data input or the nested DO loops described above. This routine then became the place where ELDIAM, frequency (FREQ), and element spacing (ELSPACE) from the reflector were defined at each run's start. In the version provided, this routine sets the design parameters for the NBS family of Yagis. Experience indicates that most of the errors leading to erroneous output from the iterations can be traced to typos made in this routine. The importance of "sanity checks" on how a Yagi's design approach is described to the model cannot be overstated. Until experience is gained in working with the model, the user should consider printing the design parameters before the iterations begin.

IMPEDANCE calculates the self and mutual impedances among the Yagi's elements. The self-resistance of all elements is assumed to be 73 ohms, and the self-reactances are calculated from Lawson's approximations. The self-impedance of the driven element needs to be 73 + j0 for best results. While the third article in this series was being prepared, it became obvious that this constant could be assigned to the driven element, eliminating the need for calculating its zero reactance length to 0.000001 inch. Hence the second element's self-impedance is set to 73 + j0 ohms. If another element is to serve as the driven element, the user has to change the test made on index J. Mutual impedances are calculated in the manner stated by Lawson, using the Kraus formulas.[3] Instead of Kraus' power series or Lawson's less-than accurate approximations, the Sine and Cosine integral routines from the IONCAP program are used. They are also three orders of magnitude faster than either of the other methods. *IMPEDANCE* also makes good use of the fact that the mutual impedance between elements i and j is the same as between elements j and i, and calculates this value only once while storing it in both matrix locations. DO loop indices I and J have an unusual implicit relationship that allows for reducing the number of calculations in a manner proportional to the number of Yagi elements.

SOLVE provides the complex current solutions for each Yagi element. It is basically a routine that solves simultaneous linear equations. An array with one more column than number of rows is necessary because this extra column contains the solved complex current values. The second element is assumed to be the driven element and is driven with 10 + j0 volts. If as was described before, a different element is to be the driven element (or in the case of multiple driven elements), the user has to make the appropriate modifications. It is also possible to obtain exactly the same gain and F/B results with any other driven element voltage value, provided that zero phase is used. Because a digital computer is being used, an assumption has to be made for the value of "*digital zero;*" this is the value used to test the complex matrix members for being equal to zero. This routine uses 0.00001 for this value. Smaller values may increase the number of iterations necessary for solving the matrix and not add anything measurable to the calculated gain and F/B. Up to this point the FORTRAN routines can be assumed to execute instantaneously. This was found to be true for Yagis with up to 20 elements. The two subroutines that follow consume the vast majority of the time necessary for a given Yagi's *iterations.*

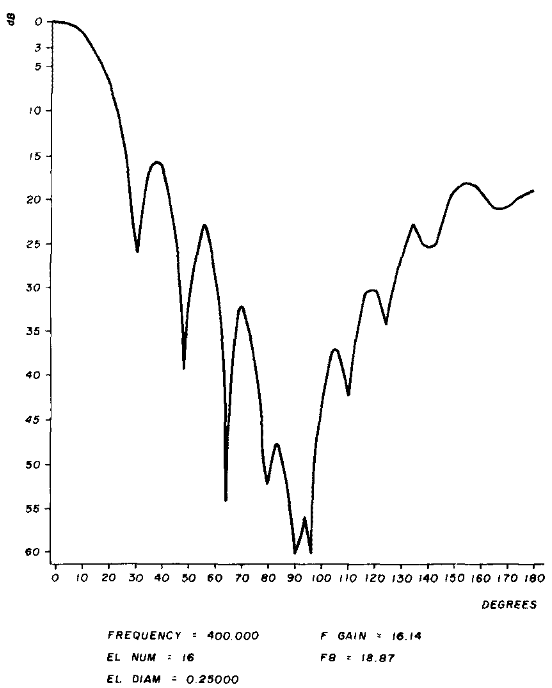

fig. 3. E-plane plot of the NBS 4.2 wavelength Yagi with an additional 0.424 wavelength-long director spaced 0.37 wavelengths from the reflector.

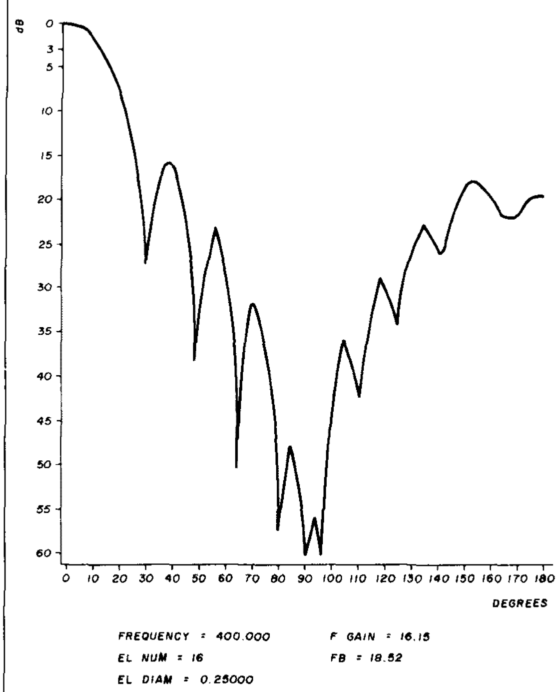

fig. 4. E-plane plot of the NBS 4.2 wavelength Yagi with an additional 0.430 wavelength-long director spaced 0.36 wavelengths from the reflector.

They contain multiple sets of nested loops that iterate each of 180 degrees within each Yagi element, or vice versa.

GAIN converts the complex current solutions in the matrix to complex voltages whose phase is in radians. A test is made for elements with zero current because they cause ATAN2 to fail. They also represent Yagis whose performance levels are undesirable. These Yagis are labelled as such and no further processing takes place. Yagi gain is calculated as a summation of each element's contribution in the forward (zero degree) direction as well as through the Poynting vectors. These values are saved for use during pattern generation and F/B calculation.

CARTPLOT calculates the F/B and either the E-plane or H-plane pattern. These patterns are then printed as normalized Cartesian plots. While coded to print only the E-plane pattern, the H-plane parameters are still computed. The user can adjust this routine to print either plane. The 180-degree intermediate values from the gain calculations are used with the dipole pattern to produce the E or H-plane plots, and to calculate F/B. If only the gain and F/B values are wanted (PLOTKEY = 0.0), a single line of output results. If a plot is wanted with full information on the Yagi (see the plots in any of the previous parts of this series), PLOTKEY is set to any non-zero value before calling CARTPLOT. Every second degree is plotted, and dB values are from zero to 60 dB. Values in excess of 60 dB are shown as 60 dB. To reliably print this many lines per page, hands-on control of the printer is almost a necessity. The user may want to change the code so the maximum dB value is 55 dB. Nulls of this magnitude are rarely achieved in practice, and are of informational value at best. To signify "off-page" values, the user may also want to change the code to print asterisks, periods, or some other symbol when these values are calculated. Because a line printer is a discrete device in terms of being able to print only in (many) fixed positions on the page, calculated dB values are rounded to the nearest integer value. The use of a pen plotter would circumvent this problem. It is also possible to use some sort of periodicity to emphasize small changes between predetermined dB values.

The **MATH LIBRARY'S** four functions are used by IMPEDANCE to calculate the mutual impedances. The resistive component is calculated by RMUT, and the reactive component by XMUT. SI and CI are the sine integral and cosine integral routines, respectively, from IONCAP. An IF statement has been added to CI to

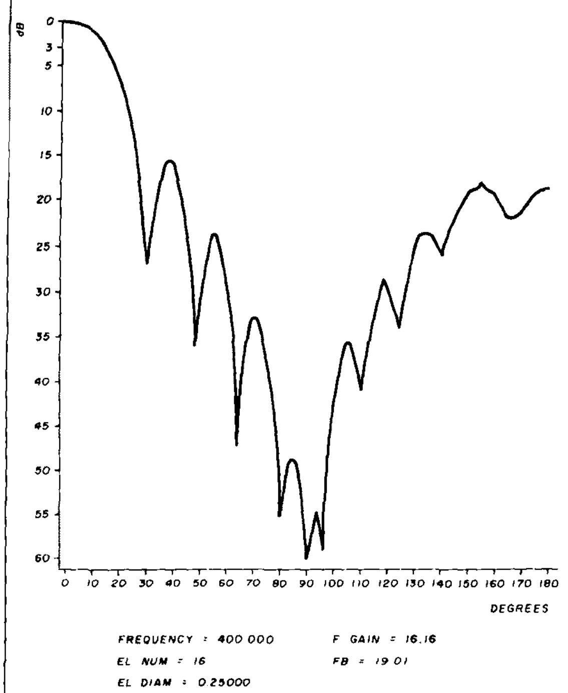

fig. 5. E-plane plot of the NBS 4.2 wavelength Yagi with an additional 0.435 wavelength-long director spaced 0.36 wavelengths from the reflector.

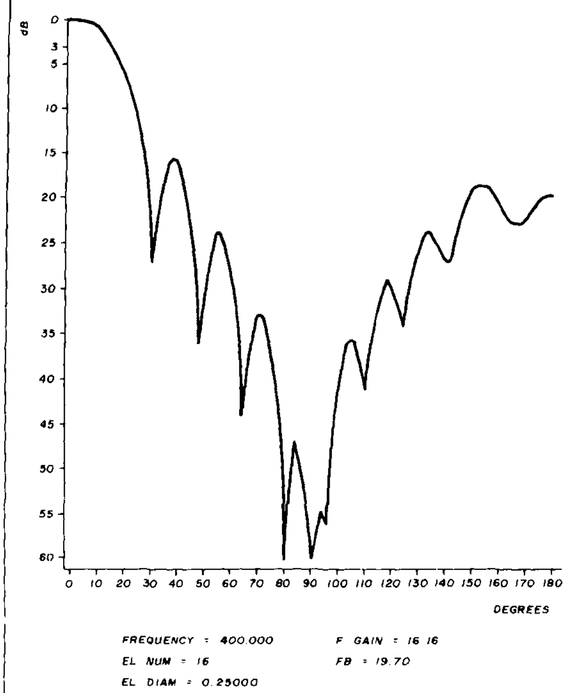

fig. 6. E-plane plot of the NBS 4.2 wavelength Yagi with an additional 0.440 wavelength-long director spaced 0.35 wavelengths from the reflector.

protect against an attempt to compute the LOG of zero. This will cause a software failure, and a small value is substituted for the zero. In actual practice a zero value cannot occur, but a simple precaution is easily taken. As with any program of this complexity, an example is in order. The 4.2 wavelength (15 element) NBS Yagi will serve as the teaching tool. In this example the model will be used to apply a long Yagi technique to this Yagi in an attempt to increase the calculated gain.

further optimizing the NBS 4.2 wavelength Yagi

Long Yagi techniques include placing one or more directors between the driven element and the first of the wide spaced directors. Up to this point the first of the wide spaced directors was the first director. Depending on the design approach used to implement long Yagi techniques, this same director may become the second, third, or even the fourth director. The Tilton/Greenblum Yagis, and to a greater extent the Knadle and Kmosko-Johnson Yagis that were examined earlier in this series, typify what is meant by long Yagi techniques. This new director (or directors) is generally believed to increase the Yagi's gain on what is substantially the same boomlength as before. This belief was shown to be true as a result of comparing computer-iterated Yagis that employed long Yagi techniques with the equal director spaced NBS Yagis of similar boom length.[4,5,6,7]

To accommodate an extra director and the spacing iterations to be done only for this director, the **DRIVER** code has to be slightly modified. The DO statement before the calls to the five subroutines is eliminated, but the associated CONTINUE is kept. GETDATA is called as before, but the NBS parameter is replaced with the integer constant 15. Code is now inserted to read the new director's length in wavelengths, and this is converted to a physical measurement at 400.0 MHz. A DO statement is now inserted to control the execution of the remaining subroutines. Its indices will serve to move the new director (element 16) between the driven element and element 3, the old first director. The driven element is 0.2 wavelengths from the reflector, and the old first director is 0.508 wavelengths from the reflector. A range of 0.21 to 0.45 wavelengths from the reflector for the extra director was selected and coded as follows:

```
DO 97 N=21,45
ELSPACE(16)=FLOAT(N)/100.0
```

table 1. A comparison between the original NBS 4.2 wavelength Yagi's calculated performance parameters and the gain maximas for a series of similar Yagis with a director added as indicated.

length of additional director (λ)	spacing of additional director from reflector (λ)	Gain (dBi)	F/B (dB)
—	—	15.70	19.26
0.424	0.37	16.14	18.87
0.430	0.36	16.15	19.52
0.435	0.36	16.16	19.01
0.440	0.35	16.16	19.70
0.445	0.35	16.16	18.95
0.450	0.34	16.15	19.44
0.455	0.34	16.14	18.28
0.460	0.33	16.10	18.29

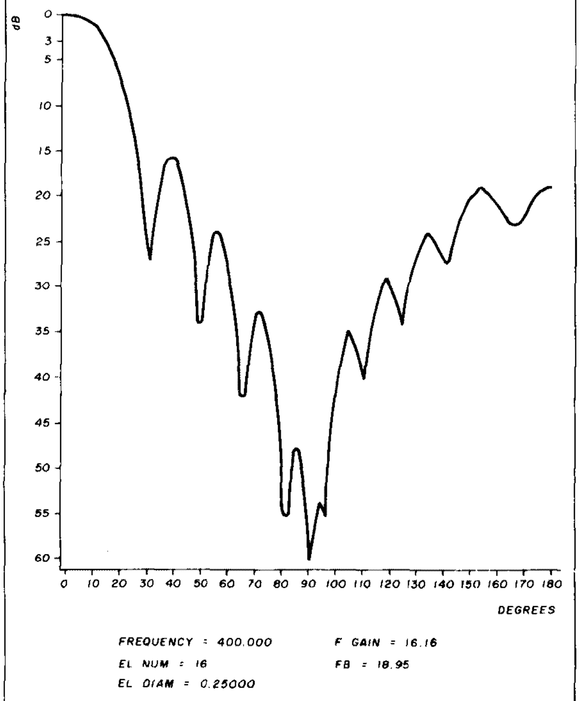

fig. 7. E-plane plot of the NBS 4.2 wavelength Yagi with an additional 0.445 wavelength-long director spaced 0.35 wavelengths from the reflector.

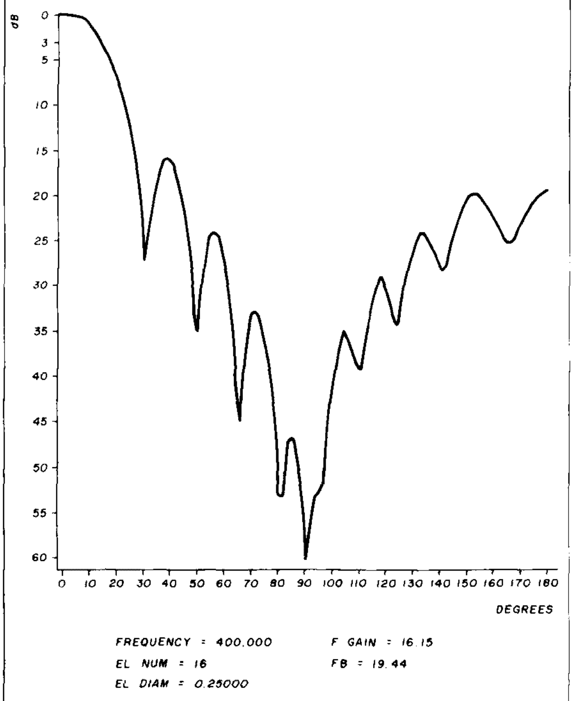

fig. 8. E-plane plot of the NBS 4.2 wavelength Yagi with an additional 0.450 wavelength-long director spaced 0.34 wavelengths from the reflector.

The remaining subroutines are called as before, but their NBS parameters are all replaced with the integer constant 16. The DO loop index should be printed during each iteration so as to identify the particulars for each line of gain and F/B calculations.

A summary of the test results of selected iterations is presented in **table 1**. Up to 0.46 dB of additional calculated gain is available by applying long Yagi techniques to this NBS Yagi. This is a significant part of the claimed 0.75 dB gain from the use of the NBS trigonal reflector, and without the additional elements on each side of the boom's plane and the attendant extra hardware.[8] **Figs. 2** through **10** contain E-plane plots for each Yagi summarized in **table 1**. While the increased forward gain is difficult to detect in the width of the main lobe, the main lobe's sharper definition through a deeper null and the reduction of many side lobe amplitudes, are visible. The calculated F/B ratios are more or less similar to that of the original Yagi and to each other. **Table 1** also indicates that as directors are moved closer to the driven element, they tend to become longer for optimal results. The new director

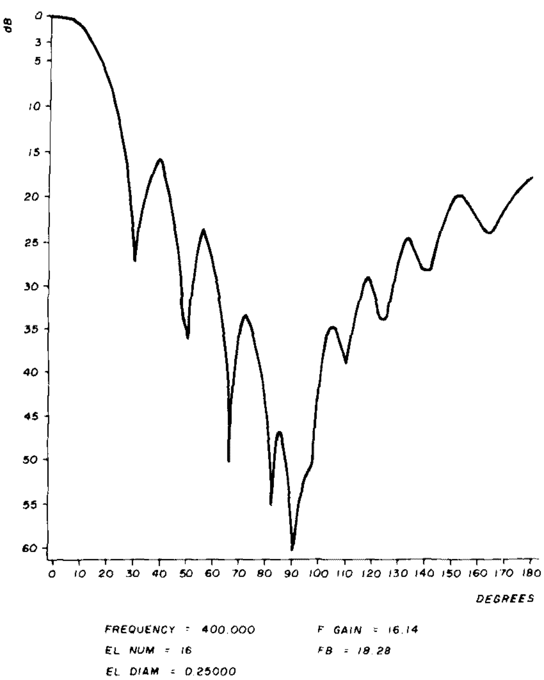

fig. 9. E-plane plot of the NBS 4.2 wavelength Yagi with an additional 0.455 wavelength-long director spaced 0.34 wavelengths from the reflector.

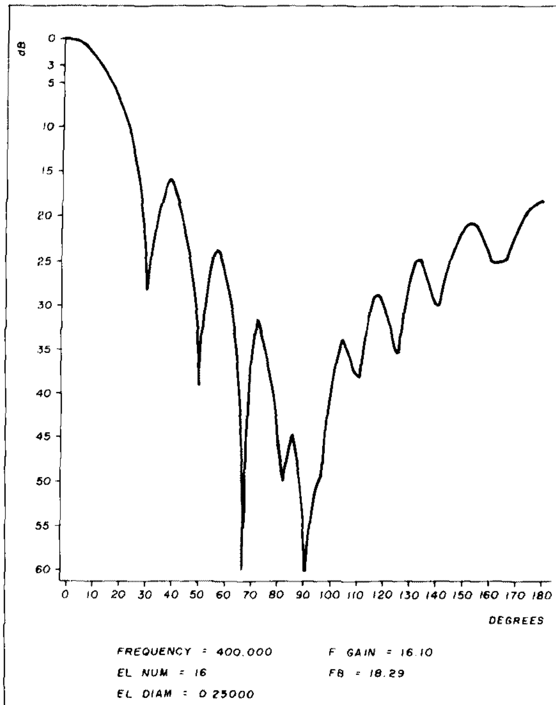

fig. 10. E-plane plot of the NBS 4.2 wavelength Yagi with an additional 0.460 wavelength-long director spaced 0.34 wavelengths from the reflector.

lengths used in these iterations begin with the length for the first director specified by NBS, and proceed in regular increments. Beyond 0.46 wavelength, Yagi performance deteriorates rapidly. With only one exception, the application of this long Yagi technique to the other NBS Yagis resulted in reduced gain. That exception produced an increase of 0.02 dB. The more experimentally inclined reader may want to iterate the model for the 4.2 wavelength Yagi with a second or even a third additional director.

conclusion

This six-part series has applied and made available a computer model for Yagi antenna design and analysis. Various Yagi design approaches were explored in terms of gain, F/B, and overall pattern, with preferable designs being selected for each band from 50 MHz to 432 MHz. The antenna is the most important part of a VHF/UHF station, and the careful selection of an antenna that is also well matched to a low-loss feedline is the single greatest step a Radio Amateur can take toward effective VHF/UHF communications. The Yagi antenna remains the most effective antenna in terms of achievable gain and minimum size. While the cost of transceivers, transverters, amplifiers, and pre-amplifiers continues to increase, very effective long Yagis can be constructed for comparatively little. This series has served to enable the VHF/UHF Radio Amateur to design that antenna with reasonable precision and with an accurate indication of how well that antenna will perform. VHF/UHF weak signal operators are basically experimenters, and it is with great interest that I look forward to future issues of *ham radio* to read the results of the experimental Yagi designs that other Amateurs produce with this model.

references

1. James P. Lawson, W2PV, "Yagi Antenna Design: Performance Calculations", *ham radio*, January, 1980, page 22.
2. James P. Lawson, W2PV, "Yagi Antenna Design: Experiments Confirm Computer Analysis", *ham radio*, February, 1980, page 19.
3. John D. Kraus, W8JK, *Antennas*, McGraw-Hill Book Company, Inc., New York, 1950.
4. Stanley Jaffin, WB3BGU, "Applied Yagi Antenna Design, part 1: A 2-Meter Classic Revisited", *ham radio*, May, 1984, page 14.
5. Stanley Jaffin, WB3BGU, "Applied Yagi Antenna Design, part 2: 220 MHz and the Greenblum Design", *ham radio*, June, 1984, page 33.
6. Stanley Jaffin, WB3BGU, "Applied Yagi Antenna Design, part 3: 432 MHz with Knadle and Tilton," *ham radio*, July, 1984, page 73.
7. Stanley Jaffin, WB3BGU, "Applied Yagi Antenna Design, part 4: 50 MHz as a Tilton/Greenblum Finale," *ham radio*, August, 1984, page 103.
8. Peter Viezbicke, WØNXB, "Yagi Antenna Design," *NBS Technical Note 688*, U.S. Department of Commerce, Washington, D.C., 1976, page 2.

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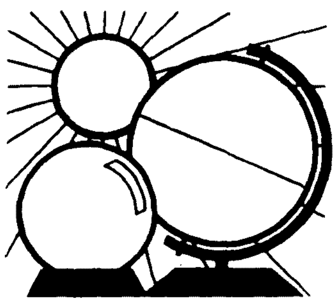

DX FORECASTER

Garth Stonehocker, KØRYW

solar cycle review/preview

A review of the past six months and a preview of the next few months of the 11-year sunspot cycle (SSN) of solar activity should be useful in forecasting propagation conditions.

The earth's geomagentic field is affected by particle influx from the sun, which in turn depends on the level of solar activity. The sun's output, expressed in flux or SSN, affects both the amount of signal absorption (and consequently the LUF, or lowest usable frequency) and the highest band we can use (the MUF or maximum usable frequency). There is a relationship between variations in signal levels (QSB) and this particle influx.

How will the signal parameters LUF, MUF, and QSB change in relation to the solar cycle by next spring? The solar flux decreased from a maximum last spring to a minimum value in August — not an unusual solar trend for most years. This year's maximum, however, was larger than most annual increases that normally occur in the winter-spring period. In addition, this spring an inflection point (increase) occurred in the recorded values of SSN two thirds of the way down from the maximum. The coincidence of the two produced an SSN level of 80 (flux 128) for the period, (February through May, 1984) about 20 SSN's above normal. This meant that MUF's were about 6 to 7 MHz higher than usual on mid-latitude propagation paths over the four-month period. Now at approximately SSN 32 we should be edging up so that next spring another slight maximum of about 8 to 10 SSNs above the average base value of 36 (flux 87) will be realized. We will check on our SSN cycle status next spring.

What does this mean in terms of working DX over the next few months? The signal strength is expected to increase because of the seasonal affect during winter as the "sun moves south," and the Northern hemisphere turns away from the sun. Not all of the signal increase is due to seasonal changes but also from the annual solar flux increase (the sun is closer to the earth) as well. At the same time, higher MUFs should provide more and longer openings on the higher frequency bands (10 and 15 meters) than at the present time. Particle influx into the polar regions are expected to be small from the minor solar flares, but may be significant from coronal hole activity, though these geomagnetic-ionospheric disturbance effects from the latter source are not intense, they are long — as long as 3 to 5 days. DX signals will probably exhibit fading (QSB) during these disturbed times. On the positive side the clouds of ionization moving around during the disturbance often focuses a weak DX signal from a new location right to you. *These enhanced signals are usually from off-great-circle propagation paths* and are short-lived in time. Be ready to take advantage of them and work them rapidly before the propagation mode quits.

last-minute update

Equinox DX conditions *should* prevail for October. Expect conditions to rapidly change during this month. As a result of fewer hours of daylight, airmass thunderstorms changing to frontal passage types, midday higher frequency band long-skip replacing short-skip, and some geomagnetic disturbances will all have their effects on the ionosphere. This month the higher frequency bands, 10 to 30 meters, will be best the first and last weeks of the month during high solar flux periods. The two weeks in between will favor the lower frequencies and nighttime DXing. Geomagnetic-ionospheric disturbances may further affect operation on October 3, 8, 13, 18, 22, and 31.

The Orionid meteor shower will be visible from the 15th to 24th of October, with a maximum rate of between ten to twenty per hour on the 20th to 21st of the month. The moon is full on the 9th and perigee occurs on the 23rd.

band-by-band summary

Ten meters will be open to the southeast for a short period before noon, to the south at noon, and to the southwest in the afternoon in local time. The openings will be longer and more frequent when the solar flux is at its 27-day cycle maximum. Even better transequatorial one-long-hop propagation will occur during disturbed periods. Tune in WWV at 18 minutes after the hour and note the geomagnetic field status announcement.

Fifteen and twenty meters, almost always open to some part of the world, will be the main daytime DX bands. Twenty should stay open on long southern paths into the night, while 15 will drop out in the late afternoon. Operate 15 first and move down to 20 meters. DX is 5000 to 7000 miles (8000 to 11,200 km) on these bands and one-long-hop transequatorial propagation is also possible, even more often than on 10 meters.

Thirty and forty meters are both day and night bands. Intermediate distances (1000 to 1500 miles or 1500 to 2200 km) in any direction represents

OCTOBER

WESTERN USA

GMT	PDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	5:00	20	20	15	10	15	10	10	20
0100	6:00	20	20	15	10	15	10	10	20
0200	7:00	20	30	20	10	15	10	10	20
0300	8:00	20	30	20	10	15	15	10	20
0400	9:00	20	30	20	10	15	15	15	20
0500	10:00	20	30	20	10	20	15	15	20
0600	11:00	20	_30_	30	15	_20_	20	15	20
0700	12:00	_20_	_30_	_20_	15	_20_	_20_	_20_	30*
0800	1:00	_20_	_30_	_30_	_15_	_20_	_20_	_20_	30
0900	2:00	_30_	_40_	_20_	_20_	_20_	_20_	_20_	_40_
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1600	9:00	_40_*	20	15	20	20	15	_20_	_20_
1700	10:00	_40_	20	15	20	20*	15	20	_20_
1800	11:00	_40_	20	15	20*	15	15	15	20
1900	12:00	40	20	15	15	15	15	15	20
2000	1:00	30	20	10	15	15	15	15	20
2100	2:00	30	20	10	15	15	10	15	20
2200	3:00	30	20	10	15	15	10	10	20
2300	4:00	20	20	15	10	15	10	10	20

MID USA

GMT	MDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN	CDT
0000	6:00	20	20	15	15	15	10	10	20	7:00
0100	7:00	20	20	15	15	15	10	10	20	8:00
0200	8:00	20	30	20	20	15	15	10	20	9:00
0300	9:00	_40_	30	20	20	15	15	15	20	10:00
0400	10:00	_40_	30	20	20	15	15	15	20	11:00
0500	11:00	_40_	30	20	_20_	20	20	15	20	12:00
0600	12:00	_40_	_30_	_20_	_20_	_20_	20	20	30	1:00
0700	1:00	_40_	_30_	_20_	_20_	_20_	20	20	_30_	2:00
0800	2:00	_30_	_40_	_20_	_20_	_20_	_20_	_20_	_30_	3:00
0900	3:00	_30_	_40_	_20_	_20_	_20_	_20_	_20_	_30_	4:00
1000	4:00	_30_	_40_	_20_	_20_	_20_	_20_	_20_	_30_	5:00
1100	5:00	_30_	_40_	_20_	20	_20_	_20_	_20_	_30_	6:00
1200	6:00	20	_30_	20	15	_20_	_20_	_20_	_40_	7:00
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2100	3:00	30	20	10	10	15	10	15	20	4:00
2200	4:00	30	20	10	15	15	10	10	20	5:00
2300	5:00	30	20	15	15	15	10	10	20	6:00

EASTERN USA

GMT	EDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE CARIBBEAN S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	8:00	20	20	15	15	15	10	10	20
0100	9:00	20	20	15	20*	15	15	10	20
0200	10:00	20	30	20	20	15	15	10	20
0300	11:00	_40_	30	20	20	15	15	15	20
0400	12:00	_40_	30	20	20	15	15	15	20
0500	1:00	_40_*	30	20	20	20	20	15	20
0600	2:00	_40_	_30_	20	_20_	20	20	20	30
0700	3:00	_40_	_40_	_20_	_20_	_20_	20	20	30
0800	4:00	_30_	_40_	_20_	_20_	_20_	20	_20_	_30_
0900	5:00	_30_	_40_	_20_	_20_	_20_	_20_	_20_	_30_
1000	6:00	_30_	_40_	_20_	_20_	_20_	_20_	_20_	_30_
1100	7:00	_20_	_30_	_20_	15	_20_	_20_	_20_	_30_
1200	8:00	_20_	30	20*	15	_20_	_20_	_20_	_40_
1300	9:00	20	30	15	15	_20_	_20_	_20_	_40_
1400	10:00	20	20	15	15	_20_	_20_	_20_	_30_
1500	11:00	20	20	15	15	_20_	_20_	_20_	_40_*
1600	12:00	20	20	15	10	20	15	20	_30_
1700	1:00	20	20	15	10	20*	15	20*	_30_
1800	2:00	20	20	10	10	15	15	15	30
1900	3:00	20	20	10	10	15	15	15	20
2000	4:00	20	20	10	10	15	10	15	20
2100	5:00	30	20	10	10	15	10	15	20
2200	6:00	30	20	10	15	15	10	10	20
2300	7:00	30	20	15	15	15	10	10	20

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during "normal" hours.

*Look at next **higher** band for possible openings.

daytime DX. Nighttime DX on these bands may be expected to offer greater distance than on 80 meters and, like 80, follow the darkness path across the sky. Reduced midday signal strengths and distances may occur on days of high solar flux values, with 30-meter openings disappearing in the pre-dawn hours on the morning after the high radio-flux value has occurred.

Eighty and one-sixty meters will exhibit short skip propagation during daylight hours and lengthen for DX at dusk. These bands follow darkness, opening to the east just before your sunset, swinging more to the south near midnight, and ending up in the Pacific areas during the hour or so before dawn. The 160-meter band opens later and ends earlier. Coastal stations and those with good low-angle radiating systems (good, long ground radials under a high vertical antenna) will usually have the edge for working rare DX. Some nights, QRN will be as low as during the wintertime DX season except when a frontal thunderstorm moves by your QTH.

ham radio

ham radio TECHNIQUES

Bill Orr, W6SAI

the beverage antenna

As old as the spark-gap and the Alexanderson alternator is the Beverage antenna. Even today it retains its popularity as a good, low-noise receiving antenna for 160 meters. Basically, the Beverage is a long wire (one to three wavelengths) suspended 10 to 15 feet above ground. Its maximum received signal level is from stations collinear with the wire (**fig. 1**). The Beverage can be made unidirectional by placing a resistive termination at the far end of the wire toward the direction of interest. Front-to-back ratios as high as 20 dB can be achieved. The value of the resistor falls between 200 and 500 ohms, depending upon the ground conductivity and the height of the wire above ground.

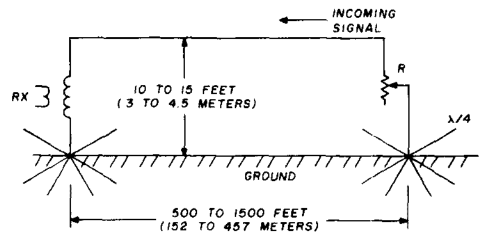


fig. 1. Single-wire, terminated Beverage antenna provides good results on 160 meters. Resistor R is adjusted for best front-to-back ratio.

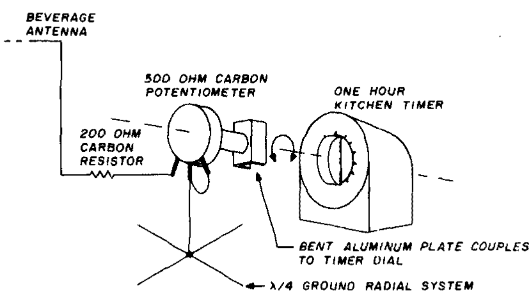


fig. 2. Simple U-plate of bent aluminum couples potentiometer to kitchen timer. Start timer and check value of front-to-back ratio of Beverage antenna over a period of an hour.

How does one determine the optimum value for this resistor? Bill Pfaff, K2GNC, has solved this problem with the device shown in **fig.** 2. His arrangement consists of a 500-ohm carbon potentiometer attached to a 1-hour kitchen timer by means of a bent aluminum coupling plate. The receiver is tuned to a broadcast station at the high end of the band and located off the back of the Beverage antenna. The experimenter now plots received signal strength against time, as measured on the timer. All that's necessary is to coordinate timer readings with the position of the potentiometer when the S-meter reading is lowest. This indicates the proper resistance (as measured on an ohmmeter) for best front-to-back ratio. The potentiometer is then replaced with a fixed resistor of the proper value.

Bill recommends that this test be run during the season of the year in which the antenna is to be used the most (probably winter), because the value of optimum resistance changes with the condition of the soil.

the K2GNC rotary ground plane antenna

Would I fool you? That's right — Bill also has a rotary ground plane. And if you don't believe me, just look at **fig. 3**. (Well, it's not *exactly* a ground plane, but I sure got your attention!)

The antenna is a modified 2-element Yagi beam working against a quarter-wave radial system. Quarter-wave elements are used as a radiator and a reflector, an idea developed by VK2ABQ. In this design, both elements are grounded to the center of the radial system and the radiator is shunt-fed by a gamma matching system. The elements are mounted in the vertical plane at an angle of about 70 degrees to each other.

Bill has made 40, 20, 15, and 10-meter versions of this simple beam. He reports it has good directivity but adds, of course, that he has no means of checking signal gain but says he gets good reports with it. The ground-mounted 40-meter version uses buried radials instead of ground-plane wires.

Standard 2-element beam dimensions, cut in half, are used in all antennas tested.

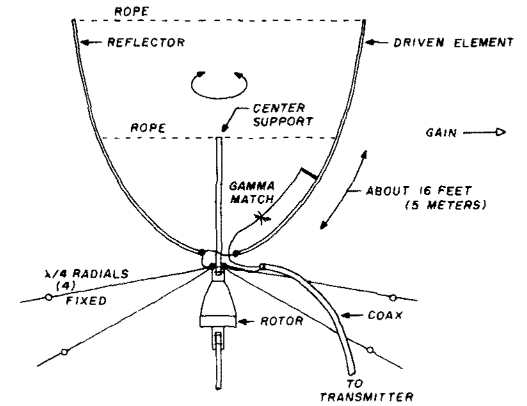


fig. 3. The K2GNC "rotary ground plane." This interesting vertically polarized array consists of a 1/4-wave driven element and reflector working against four 1/4-wave radial ground wires. The driven element is fed with a gamma match. All elements are connected to the tower (or mast) at center point, as are the radial wires.

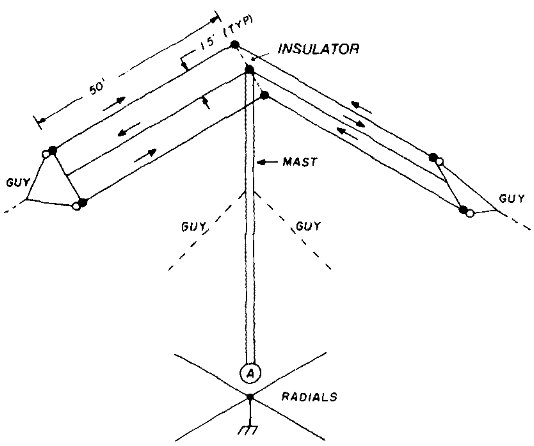


fig. 4. The K2GNC top-loaded vertical antenna for 160 meters. Flat-top currents essentially cancel and majority of radiation comes from 88-foot vertical element. Antenna is base-fed at point A via coaxial line and simple L-network. Spacing between wires is about a foot.

the K2GNC 160-meter top-loaded vertical

Still another interesting antenna concept from K2GNC is a 160-meter antenna that uses the top loading system popularized by the British Marconi Company (**fig. 4**).

The vertical antenna consists of three sections of aluminum irrigation pipe, 4, 3, and 2 inches in diameter. Its overall height is 88 feet (26.8 meters).

Bill uses 130 radials, with an average length of 100 feet (30.4 meters), and a simple matching network to feed the antenna. The antenna works, as Bill's record of WAC and 115 countries to date on the "top band" proves!

(It's always a pleasure to hear about interesting antennas; I'll be looking forward to receiving more information from readers of this column.)

the world according to JA

We're all familiar with the Great Circle Maps centered on the United States. But have you ever seen a Great Circle map with Japan as the center point? Look at **fig. 5**. Most of the countries that we consider "real DX" are within a single ionospheric hop of Japan. Such prefixes as UJ8, AP, XZ and other goodies are right in the back yard of the JA hams. On the other hand, their "tough" DX spots (on an over-the-pole shot) are such commonplace locations as VP2, HH, KV4, FY7 and the eastern tip of Brazil. Note, too, that South America runs from about 10 degrees west of North to 150 degrees south of North. And Africa is stretched out in a similar manner to the West of Japan. In Japan, as in the United States, "DX is where you find it".

SSB voice quality

I've written in the past about the poor audio quality that seems to abound in SSB communication. Must this be so? Improved response in new microphones and auxiliary speakers tend to make voices sound a bit more realistic, but some sideband signals still seem very hard to tune for good voice quality.

Maybe there's more to this than meets the eye. Intrigued by what he

heard on the bands, Willie Sayer, WA6BAN, decided to run audio measurements on some of his SSB equipment. He did this by feeding the transmitter with an audio signal generator and sweeping the audio spectrum as he measured the power output of the transmitter into a dummy load. The results of this simple test may show why some rigs sound different than others.

Fig. 6 shows the microphone-to-antenna audio characteristics of an ICOM-730 run on upper sideband. Audio response is smooth with a passband of 2128 Hz between the -6 dB points of response. The audio oscillator is swept through the range and readings logged at the points indicated on the graph.

Fig. 7 shows an old Collins KWM-1 transceiver with a Collins F-455N-20 filter in it. The curve is run on upper sideband. The response shows a characteristic filter ripple of about plus or minus 1 dB, except for a dip in the curve at about 1500 Hz. Audio passband is 1842 Hz between the -6 dB points.

Fig. 8 shows a TS-830 with a Fox-Tango filter installed. The curve is run on upper sideband. The response shows a characteristic filter ripple of about plus or minus 1dB. Audio passband is 1992 Hz between the −6dB points.

The interesting fact is that all of these transmitters sound pretty good on the air, even though the passband filters have several decibels of ripple in the audio passband. Does the IC-730 sound better on the air than the TS-830? It's hard to tell, unless exacting tests are run with the same voice, microphone, etc.

But I do know that a true audiophile would frown at a passband with even a fraction of a decibel of ripple in it!

It remains to be seen when passband filters having smoother in-band responses become available whether voice transmissions on the Amateur bands will sound more lifelike and realistic.

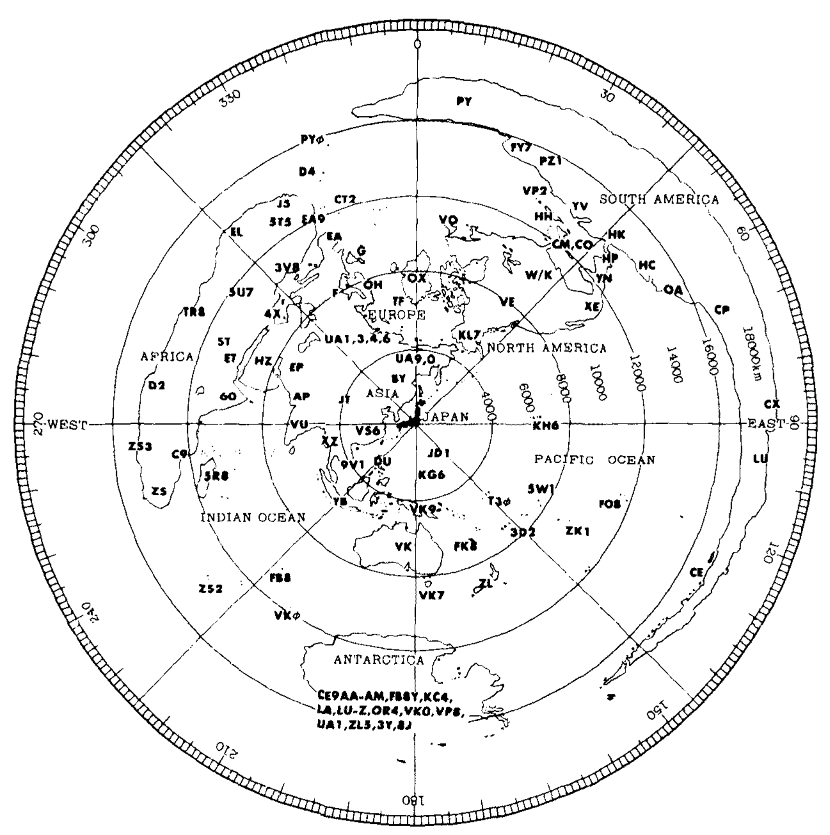


fig. 5. The DX world as viewed from Japan — Great Circle map centered on Tokyo. (Courtesy *CQ-ham radio* magazine.)

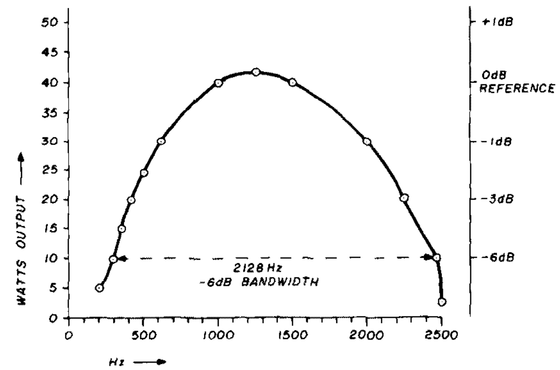


fig. 6. Passband of ICOM-730. Note smooth response with little or no cusps. Audio oscillator was swept over range of 200-2500 Hz.

It's easy to examine the audio passband of your transmitter. The test setup of **fig. 9** will do the job. Reduce transmitter power by 3 dB (half power). This lets the equipment run cooler during testing and prevents the introduction of signal distortion that would invalidate the test results.

keying a linear amplifier with the TS-930

Ron Nail, K5HUI, ran into some problems when he tried to key a linear with his TS-930 Kenwood transceiver. Here's his story:

"When I bought my Kenwood TS-930 transceiver a few years ago, I was disappointed to find that it would not operate the antenna changeover relay in my linear amplifier. Kenwood's solution was to send me a reed relay, which I installed. However, using the relay produced annoying relay clicks and also introduced a slight time lag in an otherwise excellent full break-in system incorporated in the transceiver. In addition, arcing of the antenna relay contacts was observed.

"The solution to the keying problem (shown in **fig. 10**) was to install a solid-state relay that switches in microseconds instead of milliseconds, as does the reed relay. In addition, this scheme is absolutely quiet.

"The diode bridge eliminates polarity problems on the keying line and, with the optical isolation, the circuit performs in a normal fashion, with the ability to switch up to 120 VAC or 125 VDC without regard to polarity or ground.

"A pre-packaged diode bridge (or individual diodes) can be soldered directly to the solid-state relay pins. With double-sided tape, the little unit can be attached to the chassis side panel of the TS-930, underneath the autotuner. Keying for the amplifier is pin no. 7 of the remote connector and ground."

VTR-RFI

"If you go into a contest for 48 hours and run an appreciable amount of power, you're going to meet your neighbors." That's the unhappy prediction of "Bip," W6BIP, after a de-TVI session with a video tape recorder (VTR).

Bip has spent countless hours working on his (and his neighbors') VTRs, trying to prevent his signal from entering the units. He's almost succeeded.

Because of the nature of the beast, and the lack of internal shielding, the

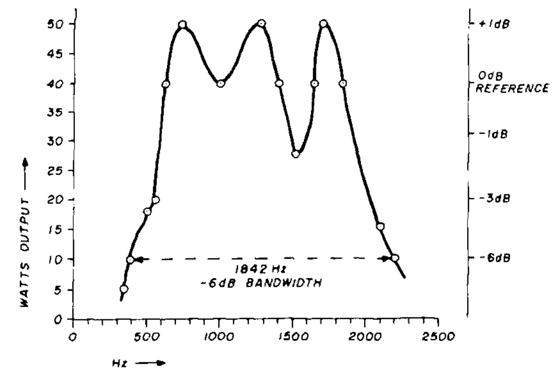


fig. 7. Passband of Collins KWM-1 transceiver with F-455N-20 filter. Note the "notch" at about 1500 Hz.

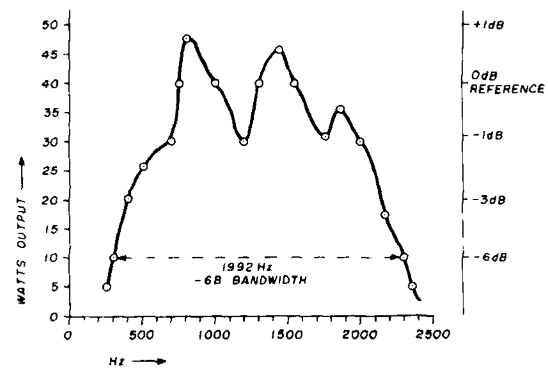


fig. 8. Passband of TS-830 with Fox-Tango filter installed.

fig. 9. Test set-up used to run audio response curves of SSB transmitter. Audio signal is inserted in microphone jack.

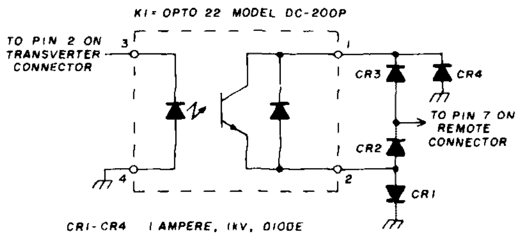


fig. 10. The TS-930 keyer of K5HUI. (The solid-state relay K1 is made by OPTO 22, 15461 Springdale Street, Huntington Beach, California 92649.)

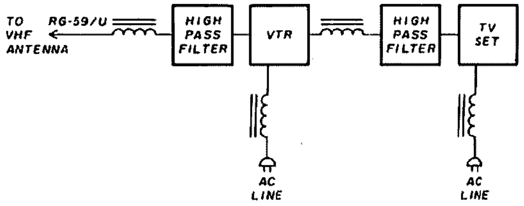


fig. 11. The W6BIP solution to VTR-RFI: wrap coaxial line to antenna and power cords to VTR and TV set around 1/2-inch diameter ferrite rods. A suitable rod is the Amidon R33-050-400 or R33-050-750. Rods have a permeability of 800 (No. 33 material).

VTR provides serious RFI (radio frequency interference) problems, particularly during playback of a tape. Interference can be expected from transmitting equipment operating on any frequency, says Bip, but the problem is worst at the low end of 80 meters, which corresponds to video processing frequencies of the unit. Transmitter power as low as 3 watts can raise havoc in the 3.5 MHz region, although it takes upward of 100 watts to cause the same picture degradation when the transmitter is operated at 3.8 MHz.*

Bip sees the challenge and rues the fact that his transmitting antenna is quite near to the TV reception antenna. But what can you do when you live on a narrow city lot?

Bip solved his RFI problem on all bands except 80 meters; on that band he has to run reduced power. But he can now run full power on all other bands up to 30 MHz using the modification of his VTR installation shown in **fig. 11.**

In this modification the TV antenna feed is RG-59/U and the line is wrapped around a ferrite rod just before it enters a highpass filter and the VTR. The power cable to the VTR is wrapped around a second ferrite rod.

The same procedures are used at the TV receiver end. The coaxial lead between the VTR and TV set is wrapped around a ferrite rod and a second highpass filter is installed at the TV set, as close to the tuner as possible. The power lead of the TV receiver is wrapped around a ferrite rod, too.

The combination of highpass filter and ferrite-wrapped coaxial lines isolates both the inner and outer conductors of the line from the strong RF field of the local transmitter. The ferrite-wrapped line cords reduce RF pickup from the power line.

That's about all you can do without digging into the VTR or otherwise modifying it to reduce RF pickup — a formidable task.

Bip points out a possible solution to RFI problems that may be available in the near future. Many problems with stereo, TV receivers, ham gear, and computers arise because the plastic cases in which the units are enclosed are "transparent" as far as RF is concerned. The equipment is completely vulnerable to any strong, nearby RF field. And in the case of the TV receiver and the computer, these devices can radiate RF "hash" that will cause problems on nearby communications receivers. Under test now is an aerosol EMI/RFI coating that can be sprayed on the inside surface of a cabinet and, when properly applied and grounded, will provide up to 75 dB of shielding effectiveness at 100 MHz. The spray adheres to any plastic, glass, or metal surface and dries to about 2 mils thickness, providing a highly conductive coating that restricts radiation through the case. This product should be available before the end of the year. While it won't solve all RFI problems, it will help the experimenter combat direct radiation and unwanted pickup by highly sensitive electronic equipment mounted within a "dust cover" that masquerades as a cabinet.†

18 and 24 MHz revisited

The Radio Society of Great Britain reports that as of January 1, 1984, 42 countries had authorized Amateur operation in the 18-MHz Amateur band. Some countries have excluded small portions of the assignment to protect local communication circuits. The main countries holding up Amateur operation in this band by their citizens are the United States and the Soviet Union.

As for the 24-MHz band, the RSGB reports that as of January 1, 1984, 42 countries had authorized Amateur operation in this band. Only Australia restricts a small segment of the band to protect a local communication circuit. Again, the main countries holding up Amateur operation in this band by their citizens are the United States and the Soviet Union.

Hello, Ivan! We have something in common at last!

*Once again this is the case of a "clean" transmitting signal being received by the VTR . . . Ed.

†Shielding can be enhanced by selection of appropriate cabinet materials and use of special EMI/RFI-resistant materials. See "Electromagnetic Interference and the Digital Era," K3PUR, September, 1984, and "EMI/RFI Shielding: New Techniques," by Vaughn Martin, January and February, 1984.

ham radio

L and C measurements

Simple circuit determines values with ease, accuracy

The typical method of measuring small value capacitors and inductors using a grid dip meter and a known value of L or C is effective, but because of circuit loading and dip meter calibration errors the results are not very accurate.

Hams who own a frequency counter can take advantage of the accuracy of this instrument to determine the value of junk box unmarked components. The technique involves measuring the frequency of an LC oscillator and then adding the unknown component to the resonant circuit, again noting the output frequency. The value of the unknown component is easily calculated using these two frequencies.

differential amplifier used as oscillator

The oscillator circuit shown in **fig. 1** uses Q1 and Q2 as a positive feedback differential amplifier. This minimizes the loading on $L1C1$. Q3, a buffer stage, isolates the counter and prevents loading down the oscillator. S1 places the unknown component in parallel with the resonant circuit $L1C1$ for capacitance measurements and in series with $L1$ for inductance measurements. In both modes the output frequency decreases in proportion to the value of the external component.

Once the effective values of $L1$ and $C1$ are known, the unknown values can be calculated from:

$$C_x = \frac{C1\,(f_1^2 - f_2^2)}{f_2^2} \qquad (1)$$

$$L_x = \frac{L1\,(f_1^2 - f_2^2)}{f_2^2} \qquad (2)$$

where f_1 is the original frequency in MHz
f_2 is the new frequency
C is in picofarads (pF)
L is in microhenries (μH)

I calculated a number of values and made up a series of graphs for inductance and capacitance values from 0.1 to 350 μH and for 3 to 700 pF. Agreement with known values is well within normal tolerance levels. My graphs are based on the initial frequency f_1 set to 9.00 MHz. Since this frequency is slightly altered by the position of S1, the core adjustment screw of $L1$ protrudes from the front panel so that f_1 can be trimmed to 9.00 MHz before checking an unknown.

As mentioned earlier, the effective value of $L1$ and $C1$ must be determined as the stray capacitance and inductance in wiring needs to be accounted for. This is easily accomplished with a known value of capacitor and inductor. Set S1 to parallel with the test probes open, adjust $L1$ to 9.00 MHz on the counter, clip on the known capacitor $C2$, and note the new frequency. The value of $C1$ is:

$$C1 = \frac{f_2^2 \times C2}{f_1^2 - f_2^2}$$

To determine the effective value of $L1$, set S1 in series with the test probes shorted and adjust $L1$ until the oscillator is at 9.00 MHz. Attach the known inductor, $L2$, to the test probes and note the new frequency, $L1$ is:

$$L1 = \frac{f_2^2 \times L2}{f_1^2 - f_2^2}$$

Use these values for $C1$ and $L1$ in formulas 1 and 2 for calculating the unknowns.

Construction is not critical; I used an L-shaped piece of aluminum to form the panel and base, and a scored piece of PC board for the oscillator circuit. A 12 inch or so length of small coax with mini-alligator clips was used to connect the unknown components.

I find I'm using this unit on every construction project. It certainly is convenient for checking the minimum and maximum capacity of small trimmers and useful in making it possible to wind a coil to a specified value — a real help when building bandpass and other types of filters.

By Larry Duthie, WB6ZLN, 1305 Lubich Drive, Mountain View, California 94040

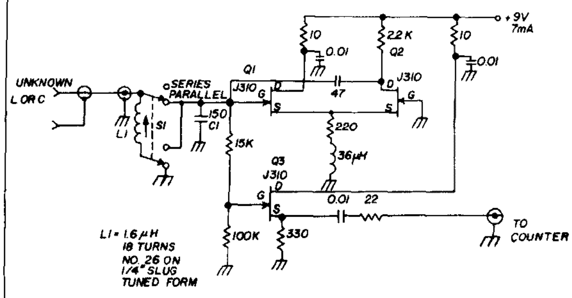


fig. 1. Oscillator circuit uses Q1 and Q2 as a positive feedback differential amplifier.

My thanks to Jim Loring, KA6VVE, for the oscillator circuit; it produces stable output over a very wide range of LC ratios.

ham radio

(continued from page 7)
asks that the 32-MHz land mobile reserved segment of the 800 MHz spectrum be released immediately for their use. — Ed.)

In summary, the 220-MHz band is not only well occupied, even in the weak-signal area, but also *necessary* to Amateur Radio, in order to handle the overflow of all types of operations from 2 meters, which would not be as well served by 70 cm; Packet Radio will also occupy some of the lower frequency regions. The gaps at the bottom are there for a purpose: to eliminate the possibility of overload that frequently hampers weak-signal operation. Weak-signal operators need this valuable portion of the spectrum as a bridge between 2 meters and 70 cm, which is vastly different in radio propagation.

— **Joe Reisert, W1JR**

*For a thorough description of all the propagation modes effective on 220 see the July and September, 1984, VHF/UHF World columns by W1JR — Editor

Smith Chart impedance matching

on your Commodore 64

Crunch the numbers, plot results with this handy 200-line program

If you have ever tried impedance matching, you've probably been introduced to the Smith Chart. Although it's fashionable to turn to computers to perform a variety of tasks, many RF types prefer to use the slower "graphic approach" because the Smith Chart provides a visual means of analysis not obvious in long columns of numbers.

But using the Smith Chart can be frustrating: *let's see — do I rotate clockwise or counterclockwise? I want a shunt element; do I follow impedance or admittance curves? Do I multiply or divide by Z_0?* Using this program **fig. 1**, the computer will do all the number crunching for you, and display your impedances on the screen within a few seconds. It will even handle errors; if you come up with an incorrect value, just press a key — your last trial *disappears*. If you like the answer, just press another key: the computer then stores that value and you can add another section.

This article assumes that you understand Smith Charts. (If you don't, see the references listed at the end of this article for an explanation.) But even if you don't understand the plot, you can still use this program because all impedances are first printed in tabular form; the plot is an option you can ignore.

program description

This program is written for a Commodore 64 using Simons' BASIC for efficient utilization of graphics commands. If you have a different computer system you can use the equations in lines **1000 through 7504** and add your own printing/plotting commands. A summary of the major program sections is listed in **table 1**, and a detailed description follows.

Lines 1 through 44 are used for program initialization. In **lines 10-11** you select a smooth or discrete plot. Later, when impedances are plotted on the screen, either selection will print a " + " on the Smith Chart at each frequency; if you select a smooth plot, a line will connect the plots to form a continuous

```
1 SP$="            ":FORJ=1TO23:CL$=CL$+SP$:NEXT
2 PRINT"<CLR><DWN><DWN><BLU>":CENTRE<ROF>"PROGRAM FOR IMPEDA
NCE MATCHING":PRINT"<C=5><DWN><DWN>"
4 CENTRE<ROF>"USING":PRINT
6 PRINT"<DWN><DWN><RED>":CENTRE<ROF>"SMITH CHART"
8 PRINT AT(<ROF>22,20)"<C=4>BY LYNN GERIG<DWN><LFT><LFT><LFT
><LFT><LFT><LFT><LFT><LFT><LFT>WA9GFR<DWN><LFT><LFT><LFT><LF
T><LFT><LFT><LFT>MARCH, 1984":PAUSE<ROF>4
10 PRINT"<CLR>DISCRETE FREQUENCIES (D) OR SMOOTH":INPUT"PLOT
(S)";D$
11 IF D$<>"D"ANDD$<>"S"THEN10
12 INPUT"<DWN>PLOT WHAT VALUE VSWR CIRCLE";VS:IF VS<1 THEN 1
2:VR=100*(VS-1)/(VS+1)
13 VR=100*(VS-1)/(VS+1)
14 INPUT"WHAT CHARACTERISTIC IMPEDANCE";ZO
15 INPUT"<DWN>HOW MANY FREQUENCIES (1-10)";N
16 FOR J=1 TO N
18 PRINT"<DWN>INPUT FREQUENCY"J;:INPUT"IN MHZ";F(J)
20 PRINT"INPUT RS, XS  OF LOAD AT "F(J)"MHZ":INPUT R(J),I(J)
22 X(J)=R(J):Y(J)=I(J):NEXT
24 PRINT"<CLR> THESE WERE YOUR LOAD IMPEDANCE INPUTS<DWN>"
26 PRINT"<DWN>      FREQ          RS          XS":PRINT
27 FOR J=1 TO N: F$=STR$(F(J))
28 RS$=STR$(R(J)):XS$=STR$(I(J))
29 USE<ROF>"   #####.###",F$:USE<ROF>"   #####.##",RS$:USE<R
OF>"   #####.##",XS$:PRINT:NEXT
30 PRINT AT(<ROF>0,18)""
32 INPUT"ARE YOU SATISFIED (Y=YES)";A$
34 IF A$<>"Y" THEN GOTO 14
39 INPUT"PRINT LOAD VALUES (Y=YES)";A$
40 IF A$="Y" THEN PRINT AT(<ROF>0,18)CL$:HRDCPY<ROF>
42 INPUT"PLOT LOAD IMPEDANCE (Y=YES)";P$
44 XM=1.4:IF P$="Y" THEN GOSUB 8000
50 PRINT"<CLR>    CHOOSE TYPE OF MATCHING SECTION<DWN>"
52 PRINT" 1 SERIES C
54 PRINT" 2 SERIES L
56 PRINT" 3 SERIES TUNED (SERIES L-C)
58 PRINT" 4 SERIES TUNED (PARALLEL L-C)
60 PRINT" 5 SERIES TRANSMISSION LINE
62 PRINT" 6 SHUNT  C
64 PRINT" 7 SHUNT  L
66 PRINT" 8 SHUNT  TUNED (SERIES L-C)
68 PRINT" 9 SHUNT  TUNED (PARALLEL L-C)
70 PRINT"10 SHUNT  TRANSMISSION LINE
72 PRINT"11 TRANSFORMER
74 PRINT"12 SERIES R
76 PRINT"13 SHUNT  R<DWN>
78 PRINT"14 STOP ADDING SECTIONS
80 INPUT"<RED><DWN><DWN><DWN><DWN>CHOICE (1-14)";M:PRINT"<C=
4>"
81 IF M<1 OR M>14 THEN 80
82 CGOTO<ROF> 500*(M+1)
```

fig. 1A. Computer aided Smith Chart design program listing, lines 1-82.

By Lynn A. Gerig, WA9GFR, RR #1, Monroeville, Indiana 46773

```
1000 PRINT CHR$(147)"ADD SERIES CAPACITOR
1002 INPUT"WHAT IS VALUE (IN PF)";C
1010 FOR J=1TO N
1015 X(J)=R(J)
1020 Y(J)=I(J)-1/(2*π*F(J)*C*1E-06)
1025 NEXT: GOTO 9000
1500 PRINT CHR$(147)"ADD SERIES INDUCTOR
1502 INPUT"WHAT IS VALUE (IN UH)";L
1510 FOR J=1 TO N:X(J)=R(J)
1520 Y(J)=I(J)+2*π*F(J)*L:NEXT:GOTO9000
2000 PRINT CHR$(147)"ADD SERIES TUNED (SERIES L-C)
2002 INPUT"WHAT IS VALUE OF C (IN PF)";C
2004 INPUT"WHAT IS VALUE OF L (IN UH)";L
2010 FOR J=1 TO N: X(J)=R(J)
2020 Y(J)=I(J)+2*π*F(J)*L-1/(2*π*F(J)*C*1E-06):NEXT:GOTO9000
2500 PRINT CHR$(147)"ADD SERIES TUNED (PARALLEL L-C)
2502 INPUT"WHAT IS VALUE OF C (IN PF)";C
2504 INPUT"WHAT IS VALUE OF L (IN UH)";L
2510 FOR J=1 TO N: X(J)=R(J)
2520 Y(J)=I(J)+(2*π*F(J)*L)/(1-((2*π*F(J))^2)*L*C*1E-06):NEX
T:GOTO9000
3000 PRINT CHR$(147)"ADD SERIES TRANSMISSION LINE
3002 INPUT"WHAT IS LINE IMPEDANCE (OHMS)";Z1
3004 INPUT"WHAT IS VELOCITY FACTOR";V
3006 INPUT"WHAT IS LENGTH (IN INCHES)";LL
3010 FOR J=1 TO N
3015 T=1.2*LL*F(J)/39.37/V
3020 D=(R(J)+Z1)^2+I(J)^2
3025 R=(R(J)^2-Z1^2+I(J)^2)/D
3030 I=2*Z1*I(J)/D
3035 Z=SQR(R*R+I*I)
3040 T=180/π*ATN(I/(R+1E-30))-2*T+180*(R<0)
3045 R=Z*COS(T*π/180)
3050 I=Z*SIN(T*π/180)
3055 D=(1-R)^2+I^2
3060 X(J)=Z1*(1-R^2-I^2)/D
3065 Y(J)=2*Z1*I/D
3070 NEXT:GOTO9000
3500 PRINT CHR$(147)"ADD SHUNT CAPACITOR
3502 INPUT"WHAT IS VALUE OF C (IN PF)";C
3506 FOR J=1 TO N: W=2*π*C*1E-06
3508 D=(1-W*F(J)*I(J))^2+(R(J)*W*F(J))^2
3510 X(J)=R(J)/D
3512 Y(J)=(I(J)*(1-W*F(J)*I(J))-R(J)^2*W*F(J))/D:NEXT:GOTO90
00
4000 PRINT CHR$(147)"ADD SHUNT INDUCTOR
4002 INPUT"WHAT IS VALUE OF L (IN UH)";L
4015 FOR J=1 TO N: W=2*π*F(J)*L
4025 D=R(J)^2+(I(J)+W)^2
4030 X(J)=R(J)*W^2/D
4035 Y(J)=W*(R(J)^2+I(J)^2+W*I(J))/D
4040 NEXT:GOTO9000
4500 PRINT CHR$(147)"ADD SHUNT TUNED (SERIES L-C)
4502 INPUT"WHAT IS VALUE OF C (IN PF)";C
4504 INPUT"WHAT IS VALUE OF L (IN UH)";L
4510 FOR J=1 TO N
4515 W=2*π*F(J)*L-(1E+06)/(2*π*F(J)*C)
4517 D=R(J)^2+(I(J)+W)^2
4520 X(J)=R(J)*W^2/D
4525 Y(J)=W*(R(J)^2+I(J)^2+W*I(J))/D
4530 NEXT J:GOTO 9000
5000 PRINT CHR$(147)"ADD SHUNT TUNED (PARALLEL L-C)
5002 INPUT"WHAT IS VALUE OF C (IN PF)";C
5004 INPUT"WHAT IS VALUE OF L (IN UH)";L
5010 FOR J=1TO N
5015 W=(2*π*F(J)*L)/(1-((2*π*F(J))^2)*L*C*1E-06)
5020 D=R(J)^2+(I(J)+W)^2
5025 X(J)=R(J)*W^2/D
5030 Y(J)=W*(R(J)^2+I(J)^2+W*I(J))/D
5035 NEXT:GOTO9000
5500 PRINT CHR$(147)"ADD SHUNT TRANSMISSION LINE
5502 INPUT"WHAT IS LINE IMPEDANCE (OHMS)";Z1
5504 INPUT"WHAT IS LINE VELOCITY FACTOR";V
5506 INPUT"WHAT IS LENGTH (IN INCHES)";LL
5508 PRINT"OPEN (O) OR SHORTED (S) STUB";
5510 INPUT S$
5512 IF S$<>"O" AND S$<>"S" THEN 5510
5520 FOR J=1 TO N
5525 T=LL*F(J)*1.2/39.37/V
5530 IF S$="S" THEN W=Z1*TAN(T*π/180)
5535 IF S$="O" THEN W=Z1*TAN((T+90)*π/180)
5540 D=R(J)^2+(I(J)+W)^2
5545 X(J)=R(J)*W^2/D
5550 Y(J)=W*(R(J)^2+I(J)^2+W*I(J))/D
5555 NEXT:GOTO9000
6000 PRINT CHR$(147)"ADD TRANSFORMER
6002 INPUT"STEP UP OR DOWN (U OR D)";T$
6004 IF T$<>"U"AND T$<>"D" THEN 6000
6006 INPUT"WHAT TRANSFORMATION RATIO";W
6010 IF T$="D" THEN W=1/W
6020 FOR J=1 TO N: X(J)=W*R(J)
6025 Y(J)=W*I(J): NEXT: GOTO 9000
6500 PRINT CHR$(147)"ADD SERIES RESISTOR
6502 INPUT"WHAT VALUE OF R";RS
6506 FOR J=1 TO N: X(J)=R(J)+RS
6508 Y(J)=I(J):NEXT: GOTO 9000
7000 PRINT CHR$(147)"ADD SHUNT RESISTOR
7002 INPUT"WHAT VALUE OF R";RS
7004 FOR J=1 TO N
7006 D=(R(J)+RS)^2+I(J)^2
7008 X(J)=RS*(R(J)^2+RS*R(J)+I(J)^2)/D
7010 Y(J)=I(J)*RS^2/D: NEXT: GOTO 9000
7500 PRINT CHR$(147);
7501 INPUT "WANT TO RUN ANOTHER ONE";A$
7502 IF LEFT$(A$,1)="Y" THEN 10
7504 PRINT AT(<ROF>14,10)"<BLU>GOOD-BY<C=4>":END
8000 HIRES<ROF> 0,0:XR=100*XM
8002 CIRCLE<ROF> 160,100,XR,100,1
8004 LINE<ROF> 160-XR,100,160+XR,100,1
8008 ARC<ROF> 160+XR,0,190,270,10,XR,100,1
8010 ARC<ROF> 160+XR,200,270,360,10,XR,100,1
8012 CIRCLE<ROF> 160+XR/2,100,XR/2,50,1
8014 CIRCLE<ROF> 160,100,VR*XM,VR,1
```

fig. 1B. Program listing, lines 1000-8014.

```
8016 CHAR<ROF> 158-.67*XR,97,43,1,1
8017 TEXT<ROF> 153-.67*XR,104,".2",1,1,8
8018 CHAR<ROF> 158-.33*XR,97,43,1,1
8019 TEXT<ROF> 153-.33*XR,104,".5",1,1,8
8020 CHAR<ROF> 156+.33*XR,97,43,1,1
8021 TEXT<ROF> 156+.33*XR,104,"2",1,1,8
8022 CHAR<ROF> 156+.67*XR,97,43,1,1
8023 TEXT<ROF> 156+.67*XR,104,"5",1,1,8
8100 CSET<ROF> 2
8105 TEXT<ROF> 10,190,"F7=RETURN",1,1,8
8110 TEXT<ROF> 246,190,"F8=PRINT",1,1,8
8115 FOR J=1 TO N
8120 D=(X(J)+ZO)^2+Y(J)^2
8125 PX(J)=((X(J)-ZO)*(X(J)+ZO)+Y(J)^2)/D*100*XM
8130 PY(J)=-2*Y(J)*ZO/D*100
8140 CHAR<ROF> 157+PX(J),97+PY(J),43,1,1
8150 IF D$="S" AND J>1 THEN LINE<ROF> 160+PX(J-1),100+PY(J-1
),160+PX(J),100+PY(J),1
8160 NEXT
8170 GET A$:IF A$="" THEN 8170
8175 IF A$=CHR$(136) THEN NRM<ROF>:RETURN
8180 IF A$<>CHR$(140) THEN 8170
8185 TEXT<ROF> 10,190,"F7=RETURN",0,1,8
8190 TEXT<ROF> 246,190,"F8=PRINT",0,1,8
8195 COPY<ROF>:NRM<ROF>:RETURN
9000 PRINT"<DWN>      FREQ          RS            XS":PRINT
9005 FOR J=1 TO N
9010 RS$=STR$(X(J))
9015 XS$=STR$(Y(J))
9020 F$=STR$(F(J))
9025 USE<ROF>"    #####.###",F$:USE<ROF>"    #####.##",RS$:USE
<ROF>"    #####.##",XS$:PRINT
9030 NEXT
9050 PRINT AT(<ROF>0,18)"F1=PLOT ON CLEAN SCREEN FOR VIEWING
9055 PRINT"F2=PLOT ON CLEAN SCREEN FOR PRINTER
9060 PRINT"F3=PLOT ON LAST SCREEN
9065 PRINT"F4=SCREEN DUMP TO PRINTER
9070 PRINT"F5=BAD VALUE:  DISCARD THIS TRY
9075 PRINT"F7=GOOD VALUE: KEEP & PROCEED
9100 GET A$:IF A$=""THEN9100
9110 IFA$=CHR$(133)THEN XM=1.4:GOSUB 8000:GOTO9050
9120 IFA$=CHR$(137)THEN XM=.833:GOSUB 8000:GOTO9050
9130 IFA$=CHR$(134)THEN GOSUB 8100:GOTO9050
9140 IFA$=CHR$(138)THEN PRINT AT(<ROF>0,18)CL$;:HRDCPY<ROF>:
GOTO 9050
9150 IFA$=CHR$(135)THEN 50
9160 IFA$=CHR$(136)THEN 9200
9170 GOTO 9100
9200 FOR J=1 TO N:R(J)=X(J)
9210 I(J)=Y(J):NEXT
9215 PRINT AT(<ROF>0,18)CL$;
9220 PRINT AT(<ROF>0,19)"<BLU>YOU NOW MUST MATCH THE IMPEDAN
CES LISTEDABOVE.<C=4>"
9230 PAUSE<ROF> 4:GOTO50
```

fig. 1C. Program listing, lines 8016-9230.

curve. The goal of most impedance matching is to find a result which is within a particular VSWR. In **lines 12-13** you select the value of a VSWR circle to appear on the plot. This makes it easy to check your results; either your plots are within the circle, or you need to refine your circuit. In **line 14** you select the characteristic impedance. This does not affect any calculations or tabular results: it is the value which is at the center of your Smith Chart plot. All impedances are normalized to this before being plotted (**lines 8120-8130**).

You are next asked how many frequencies you wish to work with (**line 15**); you then must input each frequency and the resistive and reactive series components of the starting (load) impedance at each frequency.

After you input all frequencies and load impedances, you are given a chance to start over if you made an error. Then you can choose to print the tabulated load impedances on your printer, plot them on a Smith Chart, or begin matching.

Lines 50-82 present a menu for the type of matching element to select. The program next branches to the correct set of equations for that choice (**lines 1000-7504**), and you are then asked to input those component values. For a detailed explanation of each matching section, see the section on matching network equations in the appendix.

```
THESE WERE YOUR LOAD IMPEDANCE INPUTS

     FREQ            RS           XS

       1.8           7.        -400.

ADD SERIES INDUCTOR
WHAT IS VALUE (IN UH)? 36.9

      FREQ           RS           XS

        1.8           7.          17.32

ADD SHUNT CAPACITOR
WHAT IS VALUE OF C (IN PF)? 4400

      FREQ           RS           XS

        1.8         49.89         -.37
```

fig. 2. The addition of two components converts an 80-meter vertical to 160 meters.

table 1. Major program sections.

line numbers	program content
1-10	Format, select discrete or smooth plot.
12-13	Select VSWR circle to plot.
14	Select characteristic impedance.
15-44	Input resistive & reactive parts of load impedance at each frequency. Print or plot result if desired.
50-82	Select type of matching section to try.
1000-7504	Calculation of impedances according to type of matching section chosen.
8000-8195	Subroutine for graphics plotting of Smith Chart.
9000-9170	Print new impedances in tabular form, then choose whether to print, plot, discard, or keep trial value.
9200-9230	"Good" trial value becomes new impedance to match with next section.

Lines 8000-8195 contain the graphics commands for plotting a Smith Chart on the screen. Most of these lines use graphics commands peculiar to Simons' BASIC: you may choose to write your own plotting subroutine if you have another system. A new chart is drawn in lines **8000-8023**. If you choose to plot on an old screen, you enter at **line 8100**, which recalls the last graphics screen from memory. The impedance plotting is performed in **lines 8115-8160**. The equations in **8120-8130** convert the impedances to a point on the chart normalized to the previously selected Z_0. The chart y-radius is 100 bits, and the x-radius depends on the value of "XM" (see section on printer interfaces below). **Line 8150** will connect the " + " plots with a line if "smooth plot" was previously selected. The plotted chart will stay on the screen until either function key f7 or **f8** is pressed. Pressing **f8** will copy your plot to your printer, and f7 will return you to the tabular listing of frequencies and impedances with another menu.

```
THESE WERE YOUR LOAD IMPEDANCE INPUTS

    FREQ            RS            XS

     7.             52.          -110.
     7.1            57.          -100.
     7.2            65.           -90.
     7.3            70.           -80.

ADD SERIES TRANSMISSION LINE
WHAT IS LINE IMPEDANCE (OHMS)? 50
WHAT IS VELOCITY FACTOR? .67
WHAT IS LENGTH (IN INCHES)? 130

      FREQ           RS            XS

       7.             9.74        -25.46
       7.1           11.60        -23.89
       7.2           14.15        -22.96
       7.3           16.49        -21.89

ADD SHUNT TRANSMISSION LINE
WHAT IS LINE IMPEDANCE (OHMS)? 50
WHAT IS LINE VELOCITY FACTOR? .67
WHAT IS LENGTH (IN INCHES)? 100
OPEN (O) OR SHORTED (S) STUB? S

      FREQ           RS            XS

       7.            74.46        -11.64
       7.1           59.70         -8.10
       7.2           51.39         -1.24
       7.3           45.36          2.91
```

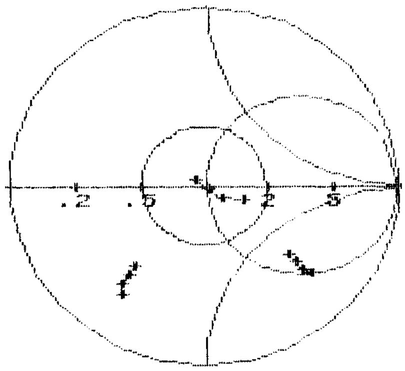
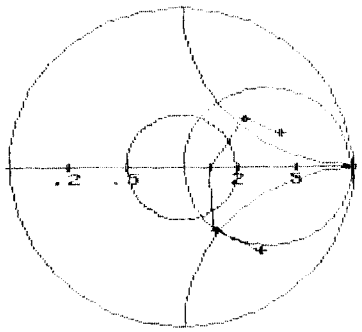


fig. 3. Compensating for tower effect on 40-meter slopers (example 2).

No matter what type of matching section is chosen, after the appropriate impedance calculations are per-

THESE WERE YOUR LOAD IMPEDANCE INPUTS

FREQ	RS	XS
110.	45.	-90.
120.	50.	-50.
130.	70.	0.
146.	80.	60.
160.	140.	90.



ADD SHUNT INDUCTOR
WHAT IS VALUE OF L (IN UH)? .06

FREQ	RS	XS
110.	17.66	60.52
120.	40.56	49.10
130.	23.02	32.88
146.	12.34	37.29
160.	12.07	47.35

ADD SERIES CAPACITOR
WHAT IS VALUE (IN PF)? 60

FREQ	RS	XS
110.	17.66	36.40
120.	40.56	26.99
130.	23.02	12.48
146.	12.34	19.12
160.	12.07	30.77

ADD SHUNT CAPACITOR
WHAT IS VALUE OF C (IN PF)? 30

FREQ	RS	XS
110.	90.94	12.62
120.	40.83	-26.88
130.	28.77	-5.40
146.	36.32	14.31
160.	87.67	-15.88

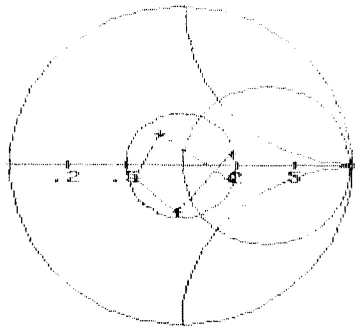


fig. 4. Broadbanding a 2-meter whip (example 3).

formed the program branches to **line 9000** and the new impedances at each frequency are printed on the screen below the values of the network components you chose. A menu appears at the bottom of the screen. You must then press a function key to continue. Pressing f1 or f2 will cause the impedances listed to be plotted on a new "clean" screen (see section on printer interfaces below). Pressing f3 will add the new plot to a previous chart. If you press f4, the tabulated data at the top of the screen is printed on your printer. Pressing f5 causes the computer to erase the bad value you just tried, and f7 causes the trial value to become the next load impedance to be matched (**lines 9200-9230**): you then branch back to the master menu at **line 50**.

examples

The three examples given below demonstrate some of the possibilities available in the program.

Example 1. My 80-meter vertical antenna is electrically "short" at 160 meters and exhibits an impedance of about 7-j400 ohms. What values of L and C do I need to match it to 50 ohms? I know I should start with a series inductor with a reactance of greater than 400 ohms (about 35 μH) then add a shunt capacitor. After experimenting with several values, I ended up with those in **fig. 2.** I only printed tabular results without graphics plot.

Example 2. When I erected my 60-foot tower, I fed the top set of guy wires as 40-meter slopers. Probably because of coupling to the tower, they were not resonant, as was to be expected. The impedances as measured at the shack end of the coax feedline are listed and plotted in **fig. 3.** Although I could have obtained a good match with a series inductor, I wanted to use leftover lengths of coaxial cable for the experience of matching with lines/stubs. The resulting impedances, shown in **fig. 3**, are an example of superimposing several plots on the same chart (using function key f3). I am presently using this network built from RG-58 cables, and the measured results are within 5 percent of those calculated.

Example 3. My 2-meter whip antenna looks like 70 ohms at resonance, including ground losses, with an increasing impedance off resonance. I wish to "broadband" match it and have less than 2:1 VSWR from

110 to 160 MHz for reception of aircraft and public service bands in addition to the 2-meter ham band. The results are shown in **fig. 4**. In this case I printed intermediate tabular results with each new element, but I plotted only the original and final impedances (on separate plots).

entering the program

Enter the program as shown taking the normal precautions to SAVE it before you RUN it, so that if you make a typing error that could cause a lock-up, you'll be able to go back to the saved version without having to retype the entire program. The following is an explanation of the mnemonics printed by my interface:

<CLR> = SHIFT-CLR
<DWN>= cursor down
<LFT> = cursor left
<BLU> = CTRL-7 (blue)
<RED> = CTRL-3 (red)
<C=4> = Commodore-4 (gray 1)

In addition, my interface sometimes prints<ROF> directly following unique Simons' BASIC commands. Ignore this when you type the program. Using Simons' BASIC on the Commodore 64, high-resolution graphics commands permit a BASIC program of only 200 lines, using about 6K of RAM. However, the equations listed are valid for *any* computer or calculator.

If you don't want to type the program yourself, I'll make two verified copies for you. (Just send me a blank tape or formatted disk with a stamped, self-addressed mailer, and a check or money order for $5.00. Use the address shown at the beginning of this article.) Without Simons' BASIC, bank switching for the high-resolution screen and point-by-point plotting needs to be done with additional programming. I also have a version for the Commodore 64 that does not require Simons' BASIC, but it is over 500 lines long and therefore not practical for listing here. I will supply a copy of it, or a similar version specifically for the VIC-20, requiring 32K expansion, under the terms described above.

appendix

The convention I used for the various matching elements and impedances is shown in **fig. A1**. The equations used for each type of reactive matching element (used in program lines **1000-7504**) follow in **fig. A2**. These equations can be used on any computer or calculator.

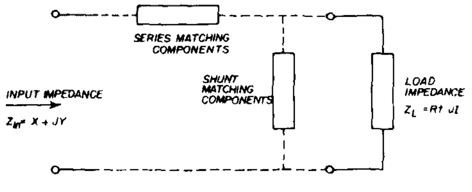


fig. A1. Matching network configurations.

impedance value to Smith Chart plot

Although the Smith Chart looks complicated, it's nothing more than a plot of the reflection coefficient, ϱ, defined as

$$\varrho = \frac{Z - Z_0}{Z + Z_0}$$

where Z_0 = the characteristic impedance you are working with and $Z = X + jY$ (your complex impedance). When plotting an impedance, the following two equations apply:

$$x\text{ - axis value} = \frac{X^2 - Z_0^2 + Y^2}{(X + Z_0)^2 + Y^2}$$

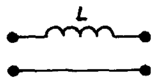
$$y\text{ - axis value} = \frac{2Y Z_0}{(X + Z_0)^2 + Y^2}$$

The value obtained will range from 0 (center of chart for $VSWR = 1$) to 1 (outside edge of chart for $VSWR = \infty$). They can be scaled by any value you desire. I multiplied by 100 for a circle with a radius of 100 dots. Equations are found in **lines 8120 - 8130**.

series L-C components

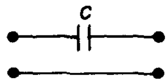
series inductor

$X = R$
$Y = I + 2\pi fL$



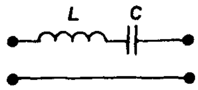
series capacitor

$X = R$
$Y = I - \frac{1}{2\pi fC}$



series L-C in series

$X = R$
$Y = 2\pi fL - \frac{1}{2\pi fC} + I$



series L-C in parallel

$X = R$
$Y = I + \frac{2\pi fL}{1 - (2\pi f)^2 LC}$

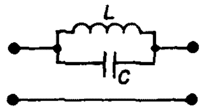


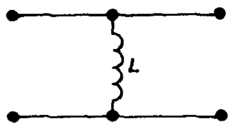
fig. A2. Equations and type of matching section used in program lines 1000-7504.

shunt inductor

$W = 2\pi fL$

$X = \frac{RW^2}{R^2 + (I + W)^2}$

$Y = \frac{W(R^2 + I^2 + WI)}{R^2 + (I + W)^2}$

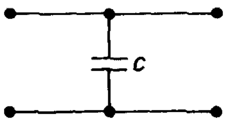


shunt capacitor

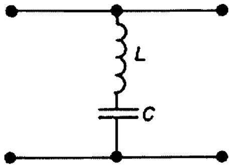
$W = \frac{-1}{2\pi fC}$

$X = \frac{RW^2}{R^2 + (I + W)^2}$

$Y = \frac{W(R^2 + I^2 + WI)}{R^2 + (I + W)^2}$



shunt L-C in series

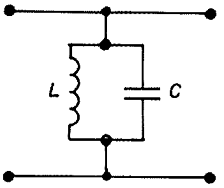


$$W = 2\pi fL - \frac{1}{2\pi fC}$$

$$X = \frac{RW^2}{R^2 + (I + W)^2}$$

$$Y = \frac{W\,(R^2 + I^2 + WI)}{R^2 + (I + W)^2}$$

shunt L-C in parallel



$$W = \frac{2\pi fL}{1 - (2\pi f)^2 LC}$$

$$X = \frac{RW^2}{R^2 + (I + W)^2}$$

$$Y = \frac{W\,(R^2 + I^2 + WI)}{R^2 + (I + W)^2}$$

series transmission line

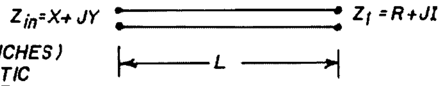
$$W = \frac{1.2\ Lf}{39.37\ V}$$

$$D = \frac{2\ Z_1\ I}{(R + Z_1)^2 + I^2}$$

$$E = \frac{R^2 - Z_1^2 + I^2}{(R + Z_1)^2 + I^2}$$

$$Z = (D^2 + E^2)^{1/2}$$

$$T = \tan^{-1}(D/E) - 2W$$



LENGTH = L(INCHES)
CHARACTERISTIC IMPEDANCE = Z_1
VELOCITY FACTOR = V

Note: If calculator gives only first and fourth quadrant for $\tan^{-1}$ answers $(-90° < W < 90°)$, add 180° to T if D was negative.

$$X = \frac{Z_1\ \{1 - (Z\cos T)^2 - (Z\sin T)^2\}}{(1 - Z\cos T)^2 + (Z\sin T)^2}$$

$$Y = \frac{2\ Z_1\ Z\sin T}{(1 - Z\cos T)^2 + (Z\sin T)^2}$$

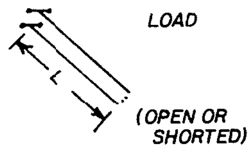
shunt transmission line (open or shorted stubs)

$$W = \frac{1.2\ Lf}{39.37\ V}$$

if open stub: $T = Z_0 \tan(W + 90°)$
if shorted stub: $T = Z_0 \tan(W)$

length = L (inches)
characteristic impedance = Z_1
velocity factor = V

length = L (inches)
characteristic impedance = Z_1
velocity factor = V



$$X = \frac{RT^2}{R^2 + (I + T)^2}$$

$$Y = \frac{W\,(R^2 + I^2 + TI)}{R^2 + (I + T)^2}$$

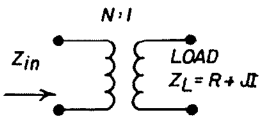
transformer

ideal transformer

N = turns ratio

$X = NR$

$Y = NI$



IDEAL TRANSFORMER

printer interface

This program was designed to calculate impedances and display them in tabular or graphic form on the screen, so you can use the program even if you don't have a printer. Just don't respond with a "yes" to questions such as "Do you want to print?"

If you do have a printer, you may want to note the two printer commands used; HRDCPY (lines **40, 9140**) gives a normal screen dump (tabular data), and COPY line **8195** copies a high-resolution (320 by 200 bits) graphics screen to the printer. Although the commands are optimized specifically for a Commodore printer, the program should run successfully with any printer/interface combination that emulates the Commodore. I use a Star Gemini-10X printer (an Epson responds the same way) with a Tymac Connection for an interface, and reproduction of the full high-resolution plot takes less than a minute. A friend with a Prowriter and Cardco interface gets similar results. Using a Cardco/+G with a Gemini or Epson printer produces the same plot, but a high-res dump takes about **45** minutes, so be patient. If you have another brand, you may need to experiment, but I am not aware of any combination that won't work.

Note: the aspect ratio (relative distance between dots on a line versus distance between lines) is different for the screen than the printer. To draw a circle on the screen, use an X/Y ratio of **1.4** (value of XM in lines **44, 9110**). This will copy as an oval on the printer. If you want a printed chart (lines **9055, 9120**), use the ratio **0.833**; this produces an oval on the screen, but a circle on the printer. If another brand of printer produces ovals, experiment with the value of XM in line **9120**. — WA9GFR

references

1. *The ARRL Antenna Book*, 14th edition, American Radio Relay League, Newington, Connecticut, 1982.
2. James R. Fisk, W1HR, "How To Use The Smith Chart," *ham radio*, March, 1978, page 92.
3. Phillip H. Smith, *Electronic Applications of the Smith Chart*, McGraw-Hill, New York, 1969.

ham radio

product

REVIEW

AEA's "Doctor DX" Morse-code contest trainer

When I visited the AEA booth *at Dayton several years ago, I sat down with Mike Lamb, AEA's president, to look at a new "mini-transceiver" they had on display. I was quite impressed with its credentials: it had all the bands I would want to work, and the sensitivity and selectivity numbers were good.*

At Mike's urging I put the headphones on and tuned around the bands to see what was going on. Twenty and 15 were hopping with DX and it seemed as if a contest were going on. Mike nudged me in the ribs. "Go ahead," he said, "give one a call." I tuned in a G3 station who was calling CQ. I dumped my call and got an immediate response: "N1ACH de G3XXX UR 599 14 BK." I went back, "G3XXX de N1ACH TNX UR 599 05 73 BK." A bell should have gone off in the back of my mind that something was amiss; there are no contests scheduled for the Dayton weekend. Oblivious to Mike's snickers, I blithely went on. After a few minutes, Mike let me in on his little secret: I was talking to a microcomputer. Needless to say I felt a little foolish. But I knew I was in good company. They'd been able to fool almost everyone who'd stopped by their booth so far. In fact, in their dozens of demonstrations, only two people figured it out.

Mike then explained what they were up to and gave me additional background on the unit. After that, I really didn't hear much about the simulator until the other day when Mike called to tell me about his newest product.

Doctor DX

Doctor DX is a cleverly designed module made up of two custom programmed EPROMs and several AEA designed integrated circuits. It fits into the cartridge port of the Commodore C-64 computer and lets you simulate chasing DX without actually being on the air. You can program Doctor DX to put your QTH anywhere around the world from Albania to Zambia. Other variable parameters are time of day, power output, and frequency of operation.

use

After installing the board into the computer and hooking up a keyer, you have to set a couple of variables before you can begin. The first is to locate your station by longitude and latitude. The second is setting the internal clock for the time of day at which you'd like to be operating. You then set your power output to one of three levels: 2, 20, or 200 watts. Finally, you select the band on which you want to operate.

Propagation is programmed to simulate what you'd hear on a given band at that time of day. For instance during the daylight hours, 10 and 15 meters will be open to DX; 160 and 80 will be dead. At night the opposite will be true: 160, 80, and 40 will be open, with the two higher bands closed. Twenty meters will be open to various parts of the world most all the time.

Another interesting aspect is that AEA has programmed the EPROMs to simulate the propagation conditions that existed during September 1979. At that time the three higher bands (20, 15, and 10) were much better than they are now. *(Gee, if you get depressed listening to 10 meters nowadays, listening to the computer will restore your faith in the band!)* This also means that propagation of the three lower bands 160, 80, and 40 is not as good as we are now experiencing but . . . you can't have everything, can you?

Callsigns are all randomly generated by the computer, with prefixes weighted by Amateur population density in that country. AEA guarantees that at least one station will be on from each of the 304 DXCC countries programmed into the computer. In an effort to replicate conditions as much as possible, stations on the lower end of each band are found working at high-speed CW (30-40 WPM). The higher you go on each band, the slower the stations are. Contacts consist of a signal report followed by CQ WW zones. If you miss an exchange, you can ask the station for a repeat. You can also request that the station speed up (QRQ) or slow down (QRS).

There are four frequency tuning rates with Doctor DX. The function keys on the right of the Commodore keyboard are used to "tune" the receiver up and down frequency. F-1 is FAST UP, to 20 kHz per second, F3 up 2 kHz per second, F5 is FAST DOWN – 20 kHz per second and F7 is down – 2 kHz per second.

So that you can follow your progress with the test, the computer is programmed to constantly display your score and QSO rate. Contest lengths can be varied from a one-hour sprint to a test of several hours' duration.

operation

Now that everything is all hooked up and location, time of day, power, and frequency (160 meters, of course) are all programmed in, it's time to get this mock contest started. The first station worked was a K2, and I even got a 599! Hey, I'm off and running!

The first thing to decide is whether I'm going to find a clear frequency and call CQ or do a "hunt and pounce" contest. Since I want to try both, I start by calling CQ. After two calls, a VE3 comes back with a 589 in zone 5. I quickly work another 10 stations before a VE2 starts calling CQ up a few kHz from mine. However, I hear a weak call, and ask QRZ? By golly, it's my first DX, an HI7. He gives me a 589 and zone 8. I shoot back his report only to have him ask twice for a repeat. Repeat? I am talking to a computer aren't I? After a couple of minutes, I'm not so sure. (Mike's done it again. I actually can believe I'm in the midst of a CQ WW!)

Now it's time to do some looking to see what I can find. I change bands and start at the bottom, looking up frequency. The first station I tune by is running at least 40 WPM. I call him twice only to have him go back to two other computer calls. Finally he calls me and I put another into the log.

After about an hour's worth of operation it's time to see how I've done. I have 28 contacts on three bands, with the majority (22) on 160 meters, 4 on 40 meters, and one TI3 on 80 meters. I'm sure I would have done better during my one-hour sprint if there'd been no phone calls or other interruptions.

My QSO rate varied from 35 to 45 QSOs per hour while I was able to dedicate full time and attention to the test. I feel quite strongly that with a little practice my rate during actual contests would improve dramatically.

There's the real beauty of Doctor DX. No matter where your interest in Amateur Radio lies, using Doctor DX will help improve your CW operating skills. After a couple of hours of use, you'll find that code copy comes much more easily. One of our own Novices stopped by and watched while I was running the test and commented that the real-life simulation offered by Doctor DX would be a tremendous help in improving CW skills and operating confidence in a way that wouldn't be embarrassing to the beginner. Experienced contesters will find operating a contest from EP2 to be quite a change from their normal QTH. I think that many will find the added insight of having worked portable EP2 to give them a tactical edge during actual contests. Doctor DX can be many things to different people. I think that those who use it will find it to be very rewarding and *plenty* of fun.

awards

AEA has gone whole hog here and is offering the following awards:

AEA-CC: for working 100 different countries
AEA-AZ: for working all CQ zones
AEA-DC: for working 100 countries on five bands
AEA-HR: AEA's honor roll for working 250 countries

Handsome certificates will be awarded for each level reached. These awards are endorsable or you add to your totals. AEA will also be keeping a running tally of high claimed scores and will periodically publish the current top five scores in two categories: Marathon (24 hours) and Sprint (8 hours). There is a minimal charge for these awards.

The price of Doctor DX will be $149.95 from AEA dealers and the unit should be available by the time you read this review. It is a perfect Holiday gift for that hard to buy for ham friend. Clubs will also find it quite useful for training in sprint contests or in any number of other applications.

For information, contact AEA, P.O. Box C2160, Lynnwood, Washington 98036.

Circle #301 on Reader Service Card.

de N1ACH

NEW products

"Remote-A-Pad"

Remote control of an HF or VHF station is now possible using the "Remote-A-Pad" from Engineering Consulting. It incorporates the radio's own digital frequency control inputs (keypad) and a simple eight wire connection to the control pad of the radio. The audio from, for example, a two meter rig is decoded and turned into rows and column switches which provide the necessary logic to work in parallel with an existing keypad.

The "tough tones" from your handheld's keypad are transmitted to another receiver which provides the audio link into the "Remote-A-Pad." The tone selected then activates the keypad. If you hit a #1 on the portable then a #1 on the radio to be controlled is activated. When controlling radios like the ICOM-701 and using the RM-2 controller the power of the microprocessor already built into the rig is unleashed. Many other radios can also be used with the "Remote-A-Pad," in fact any radio or rotator (Pro Search for example) having a 3 × 4 or 4 × 4 keypad may be controlled since the rows and columns are found on all keypads which are used to control computerized devices.

Find yourself a radio which is now controlled with a keypad and install a "Remote-A-Pad." Add a phone patch for audio link, and the "Remote-A-Pad" will also turn on and off your link with two on board 4 digit DTMF (touch tone) programmable decoders. You set the code to any series of the 16 DTMF tones. LEDs on the board will tell you the status of the controlled device (on or off). A momentary pulse is also provided and may be used directly into logic to control on-off devices. Relays may be added by using a transistor driver.

The "Remote-A-Pad" is shipped with a 22-(gold) pin card edge connector as well as a 16-pin dip socket ribbon cable for easy installation. Detailed installation instructions and schematics are provided for $149.00.

For further information, contact Engineering Consulting, 583 Candlewood Street, Brea, California 92621.

Circle #302 on Reader Service Card.

satellite converter

The Ten-Tec Model 2510 Mode B satellite station incorporates a complete 435 MHz, 10-watt output, single sideband or CW transmitter and a 2-meter to 10-meter receiving converter, both tuned with a common frequency-adjusting knob.

The transmitter is VFO tuned across the required 435.0 to 435.5 MHz segment and features selections of CW or SSB modes and of upper or lower sideband. The output power level is automatically maintained at a present level (ALC) so that overdrive is not possible. The preset level can be reduced from the factory-set 10-watt level if a linear power amplifier with drive requirements below this level is used.

The receiving converter employs a GaAs FET preamplifier for lowest noise figure and an output frequency of approximately 29.0 MHz. The 10-meter band of any Amateur HF transceiver or receiver, or a high-quality general coverage receiver, can be used in conjunction with the converter. Once the HF receiver is set to the correct frequency, *all* satellite tuning is performed with the tuning knob. (Transmitter and receiving converter are automatically tracked, thus producing a pseudo-transceiver.)

Provisions are made to allow duplex operation so your transmitted signal can be received simultaneously for system and frequency checks. A front panel SPOT pushbutton switch places the transmitter in operation with carrier inserted.

Priced at $489, the unit simplifies station assembly, reduces the number of separate items and interconnections, eliminates the need to buy separate converters, and converts your HF station into a transceiver type OSCAR station.

For additional information, contact Ten-Tec, Inc., Sevierville, Tennessee 37862.

Thruline® wattmeter elements

A new series of five plug-in elements, each of which offers bi-directional RF power measurement from 20 milliwatts to 100 watts at ±5 percent accuracy of reading, in frequency bands from 25 MHz to 1000 MHz, is available for the Bird Thruline® Wattmeter. Full rated accuracy is maintained over this 5000 to 1 power range, with temperature extremes between 0 and 50 degrees C.

The entire complement of Model 4410 elements — each with seven full-scale power levels instead of one — enables the new wattmeter to service CW and FM systems from 200 kHz to 1000 MHz and from 20 milliwatts (on the 100 mW scale) to 10,000 watts. The desired range is selected instantly by a front-mounted rotary switch. A choice of over 18 quick-change connectors mate with any common RF type.

This sensitive instrument is appropriate for any application where accurate in-line measurements at milliwatts, watts, or kilowatts need to be performed fast, economically, and simply at any skill level, independent of AC line power, in indoor our outdoor environments at freezing or tropical temperatures.

The new low power plug-in elements are priced at $125; the Model 4410 Thruline® Watt-

NEW products

meter (with Quick-Change female N connectors) is $495; other elements are priced from $125 to $175.

For additional information, contact Bird Electronic Corporation, 30303 Aurora Road, Cleveland (Solon), Ohio 44139-2794.

Circle #304 on Reader Service Card.

HF antenna accessories catalog

Microwave Filter's Catalog PC/84 describes baluns, multiband traps, and remote antenna switching relays for the 2 to 1000 MHz communications bands.

The catalog illustrates the construction of single and multiband dipoles in the 1.8 to 30 MHz Amateur and HF communication bands and offers other construction accessories such as end insulators, weatherized center-feed insulators, wire, cable, and connectors.

The system of remote, voltage-controlled antenna switching relays allows selection of up to nine separate antennas through a single coaxial cable as well as direction changing of antenna arrays and automatic pairing of antenna/transmitter sets.

For a copy, contact Microwave Filter Company, Inc., 6743 Kinne Street, East Syracuse, New York 13057.

Circle #308 on Reader Service Card.

simplex autopatch system

Hamtronics, Inc., has offered a Repeater Autopatch in kit form, including the DTMF Tone Decoder/Controller module for several years. They now have a new module that can be used with the Autopatch to allow operation on simplex transceivers. When an autopatch is used on a repeater, the duplex operation of the repeater allows the mobile operator to break into the telephone conversation anytime; when operating simplex — that is, transmitting and receiving on the same frequency — it is necessary to have a method to allow the mobile operator to access the base station receiver even though the base transmitter is on the same frequency.

The new AP-2 Simplex Autopatch Timing Module kit provides the required timing and logic circuits to allow the mobile operator to bring up the autopatch through touch-tone control, and it keys the base station transmitter with an automatic window generator circuit, which breaks periodically to allow the receiver to listen for the mobile station. If the mobile station is on the air, the window generator stops the base transmitter for the duration of the mobile transmission so the mobile operator maintains complete control of the transmit-receive switching of the base transceiver. Window length is automatically adjusted to compensate for the characteristics of the transceiver for minimal disruption of mobile reception by the window in the base transmissions. The new AP-2 module along with the basic Autopatch and DTMF Tone Decoder modules are priced at $200. The system is also available wired and tested.

For more information, contact Hamtronics, Inc., 65 Moul Road, Hilton, New York 14468-9535.

Circle #305 on Reader Service Card.

terminal interface

Terminal Interface sends messages around the world using any personal or home computer. Compatible with any home computer, such as Heath's H-8 or H-89, Commodore, Atari, Radio Shack, Apple, or IBM, it sends or receives ASCII/Baudot RTTY and Morse Code messages at up to 300 Baud using any standard transceiver, TTY terminal or monitor, and a computer with the appropriate software.

Optional six-pole filters for the HD-3030 include a preselect filter that delivers strong, readable tones in the standard 170 Hz shift and print clearly even in a crowded band, as well as filters for 425 and 850 Hz audio shifts. Features include a crystal-controlled AFSK generator, capability for full FSK with equipped transmitters, true mark/space detection, oscilloscope tuning outputs, front panel LED bargraph tuning, and data and status indicators. TTL and RS-232C I/O compatibility and a built-in loop supply are also included. A row of flag-type pushbuttons permits full up-front control of send and receive (including reverse shift) configurations. A convenient autostart relay energizes the rear panel AC receptacle for unattended start of the computer and/or printer, while an internal thershold adjustment sets the desired recognition level.

The Terminal Interface kit includes a mate for the DB-25 socket and a step-by-step assembly manual that supplies a pin-out and criss-cross interconnection chart.

Heath also offers Super CW and Super RTTY Terminal Interface software for use with their H-8 and H-89 computers.

For further details, contact Heath Company, Benton Harbor, Michigan 49022.

Circle #306 on Reader Service Card.

self-contained, transportable HF antenna system

A self-contained, transportable log-periodic HF antenna system has been announced by Telex/Hy-Gain. The system, Model 5025, consists of the antenna, rotator, 60 foot (18.3 meter) telescoping tower, generator, and installation tools and hardware, all stowed on one tandem-axle trailer. The compact design permits all

modes of transportation from towing with a pick-up truck to airlifting on a C-130 aircraft.

The antenna covers the frequency range of 4 to 30 MHz, with a gain of 9 dBi from 4 to 6 MHz, and 12 dBi from 6 to 30 MHz. The electric antenna rotator can be controlled from the base of the tower or remotely with a two-wire remote control unit which is provided. The remote control features a digital display of the antenna azimuth in 10-degree increments.

Maximum use of color coding, captive hardware, and modular construction makes the 5025 the fastest erectable rotating tower/antenna system in its class. The system can be assembled and erected by as few as two persons; a team of four can complete the task in less than two hours. The self-contained 5 kW generator set provides all power necessary for erection and operation of the antenna. A hydraulic tilt-up trolley provides positive control during erection. This feature is said to yield a high level of safety and eliminate the complexity and hazard associated with the manual erection of towers.

For additional information, contact Telex Communications, Inc., Hy-Gain CIM Department, 8601 N.E. Highway Six, Lincoln, Nebraska 68505.

microprocessor controlled repeater

The new SCR2000X microprocessor-controlled repeater from Spectrum Communications includes full autopatch and touch-tone repeater remote control capability and patch AGC for constant levels, as well as phone line and "over-the-air" command modes.

Able to store up to 13 "auto dial" phone numbers, the unit includes a touch-tone to dial pulse converter and offers full 16-digit decoding with crystal controlled decoder IC, allows the use of "A, B, C, and D" characters in control codes. This expands the number of possible codes and increases security.

Touch-tone control of all important repeater functions includes timeout, hang time, patch timeout, transmitter inhibit/reset, patch and reverse patch inhibit/reset, and "PL" ON/OFF (with optional PC board). A built-in battery back-up saves microprocessor memory in case of power failure.

The following transmitter options are available: 2 meters: 30 or 75 watts; 220 MHz: 30 or 65 watts; 440 MHz: 40 watts. High power rack mount repeater power amps and power supplies are available to 150 watts.

A high-performance receiver is included, with high sensitivity, selectivity and wide-dynamic range. An 8-pole front end filter is standard as well as a 12-pole IF filter. "Super Sharp" filter options are also available.

For further details, contact Spectrum Communications Corp., 1055 W. Germantown Parkway, Norristown, Pennsylvania 19401-9616.

Circle #310 on Reader Service Card.

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ROHN TOWERS — Wholesale direct to users. All products available. Write or call for price list. Also we are wholesale distributors for Antenna Specialists, Regency, and Hy-Gain. Hill Radio, PO Box 1405, 2503 G.E. Road, Bloomington, IL 61701-0887. (309) 663-2141.

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TRAVEL-PAK QSL KIT — Converts post cards, photos to QSLs. Stamp brings circular. Samco, Box 203-c, Wynantskill, New York 12198.

ULTIMATE CW FILTER: Four pole Gaussian response with variable selectivity and center frequency. Includes audio amplifer. See October 1983 QST, page 14. Complete PC assembly kit only $21 postpaid. Orders or information: Analog Technology, Box 8964, Fort Collins, CO 80525.

RADIO ITEMS before 1930 wanted. Buying battery operated radios, horn and cone speakers, radio tubes and parts, radio literature — books, catalogs, magazines, radio advertising signs, posters. Gary Schneider, 6848 Commonwealth Blvd., Parma Heights, Ohio 44130.

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DIGITAL AUTOMATIC DISPLAYS for FT-101's TS-520's, and most others. Six 1/2" digits. Write for information. Grand Systems, P.O. Box 2171, Blaine, Washington 98230. (604) 530-4551.

SQUIRES SANDERS, SS1R (701 series); SS1V-SS1-S, SS1-MS mint condition. Factory manuals and other parts. Heath HX-10 transmitter with accessories, HO-10 scope. Icom IC-280 FoMoXceiver, Kenwood TR-2400. All in excellent condition with manuals. (616) 382-5401. K8WPQ, 9549 No. 17, Kalamazoo, MI 49007.

RUBBER STAMPS: 3 lines $4.50 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

FOR SALE: Kenwood TS-520 still in box. Operated once to check out. Great buy $400.00. MFJ antenna tuner $15.00. Hy-Gain 18 AVT vertical antenna $20.00. Ham Radio magazines back to 1980 $1.00 each. Call Warren (617) 335-7756 nights 7 PM to 10 PM. No calls on Sundays please.

FOR SALE: Swan 350 transceiver w/ps $250. SA2040 antenna tuner $125. QF-1A audio filter $50. Kantronics CW/RTTY interface for Vic-20 with programs board + cables $135. Shipping included. Send money order. Package price $500. Write Jim Howell, KA4EBW, 18 Dan St., Salisbury, NC 28144. (704) 637-0313 evenings.

DRAKE TR7 TRANSCEIVER with two optional filters, one fan and hand mike $625. PS7 power supply $200. Will sell rf deck separately. 182 feet of new Belden RG59U coax $30. All mint. You ship. (609) 627-5683.

REPAIR, ALIGNMENT, calibration. Collins written estimates $25; non Collins $50. K1MAN. (207) 495-2215.

AN/PRC-74 AND PRC-104 transceivers, modules and accessories wanted. T. Stroh, N6JSD, 21930 Valerio St., No. 11, Canoga Park, CA 91303.

CHASSIS and cabinet kits. SASE K3IWK.

GET THE FACTS! Let our *rf Snitch fill* you in. Attach this sensing device to your meter to detect rf power. Use for signal tracing in receivers, and tuning antennae, transmitters, oscillator and rf tank circuits. For additional information write: K8MLN or N6DPW, Engineering Research Assoc., PO Box 170, Ontario, Ohio 44862.

NATIONAL RADIO CO equipment manuals price list SASE. Dust covers, NCX 3 or NCX 5 plus NCX A, pair $8.95 PP. Maximilian Fuchs, 11 Plymouth Lane, Swampscott, MA 01907.

TI 99/4A Random Code practice programs. Dr. Code "General" sends international Morse code and prints on screen; you choose: speed, tone, which characters to be sent, spacing, and more! Dr. Code "Speech" same as "General" with speech; you choose how many characters before speech check. For cassette of both copyrighted programs and conditional copying privileges, send $10.00 plus $3.00 shipping and handling to N5ESF, Rt. 1, Box 1326, Lake Charles, LA 70601.

ELECTRON TUBES: Receiving, transmitting, microwave . . . all types available. Large stock. Next day delivery most cases. Daily Electronics, 14126 Willow Lane, Westminster, CA 92683. (714) 894-1368.

S9 PLUS QTH. 59 acres for antenna farm — elevation above Derry repeater, equals Francestown repeater — south slope Crotched Mountain for wide propagation window, solar exposure and view — spacious 4-room lodge, baths, separate bunkhouse with shop and garage, excellent condition — paved access — privacy — one hour from Nashua or Manchester — $95K — must liquidate. "CW" Farr, Broker, Greenfield, NH 03047. (603) 547-2053.

WANTED: Cash paid for used speed radar equipment. Write or call: Brian R. Esterman, PO Box 8141, Northfield, Illinois 60093. (312) 251-8901.

$$$$ SUPER SAVINGS on electronics parts, components, supplies, and computer accessories. Free 40-page catalog for SASE. Get on our mailing list. BCD Electro, PO Box 830119, Richardson, TX 75083. Or call (214) 690-1102.

WANTED: Old microphones, remote mixers other misc related items. All pre-1935. Box Paquette, 107 E. National Avenue, Milwaukee, WI 53204.

PORTABLE 2-meter Quads and J-Verticals. Write Radio Engineers, 3941 Mt. Brundage Avenue, San Diego, CA 92111.

RECONDITIONED TEST EQUIPMENT $1.00 for catalog. Walter, 2697 Nickel, San Pablo, CA 94806.

FOX-TANGO Newsletters — Since 1972, the prime source of modifications, improvements, and repair of Yaesu gear, free to Club members. Calendar year dues still only $6 U.S., $9 Canada, $12 elsewhere. Includes five year cumulative index by model numbers, or send $1 for index and sample Newsletter. Fox Tango Club, Box 15944, W. Palm Beach, FL 33416.

RTTY-EXCLUSIVELY for the Amateur Teleprinter. One year $7.00. Beginners RTTY Handbook $8.00 includes journal index. P.O. Box RY, Cardiff, CA 92007.

IMRA International Mission Radio Assn. helps missioners — equipment loaned; weekday net, 14.280 MHz, 2-3 PM Eastern. Br. Frey, 1 Pryer Manor Rd., Larchmont, NY 10538.

"HAMS FOR CHRIST." Reach other Hams with a gospel tract sure to please. Clyde Stanfield, WA6HEG, 1570 N. Albright, Upland, CA 91786.

TENNATEST — Antenna noise bridge — out-performs others, accurate, costs less, satisfaction guaranteed, $41.00. Send stamp for details, W8URR, 1025 Wildwood Road, Quincy, MI 49082.

WANTED: Early Hallicrafter "Skyriders" and "Super Skyriders" with silver panels, also "Skyrider Commercial", early transmitters such as HT-1, HT-2, HT-8, and other Hallicrafter gear, parts, accessories, manuals. Chuck Dachis, WD5EOG, The Hallicrafter Collector, 4500 Russell Drive, Austin, Texas 78745.

VERY In-ter-est-ing! Next 4 issues $2. Ham Trader "Yellow Sheets", POB356, Wheaton, IL 60189.

Coming Events
ACTIVITIES
"Places to go..."

PENNSYLVANIA: The Irwin Area ARA will sponsor a Swap & Shop, Saturday, October 20, Circleville V.F.D., off Rt. 30, 3.5 miles west of Pennsylvania Turnpike, exit 7. Talk in on 325/925 and 52. For information: Don Myslewski, K3CHD, 359 McMahon Road, North Huntingdon, PA 15642. (412) 863-0570.

LOUISIANA: The Twin City Hams are sponsoring a Hamfest, Saturday, November 10, Convention Center, West Monroe. Swap tables, new dealers, exams, ladies' events. All indoors. Talk in on 146.25/85. Contact: Benson Scott, AE5V, 107 Contempo, West Monroe, LA 71291.

TENNESSEE: Hamfest Chattanooga and the Tennessee State ARRL Convention, October 27 and 28, Memorial Auditorium, Oak Street at Lindsay Avenue, Chattanooga. Forums, contests and non-ham activities. Amateur exams Saturday morning, October 27, at 8 AM, Tech through Extra. Eight foot flea market tables indoors available $6.00 per day or $10.00 both days. For information: Hamfest Chattanooga, PO Box 3377, Chattanooga, TN 37404 or call Nita Morgan, N4DON. (404) 820-2065.

KANSAS: Sandhills ARC's Swapfest, Sunday, October 7, 4-H Building, Scott County Fairgrounds, Scott City. Doors open 9 AM. Talk in on 146.10/.70. Covered dish lunch.

GEORGIA: Rome Hamfest (the South's oldest) will be held Sunday, October 1, Civic Center, Rome. Starts 8 AM. Talk in on 147.90/30. Contact: T.J. Freeman, (404) 232-2830.

OHIO: The Marion Amateur Radio Club's 10th annual "Heart of Ohio" Ham Fiesta, Sunday, October 28, 0800 to 1600, Marion County Fairgrounds Coliseum. Tickets $3.00 advance, $4.00 door. Tables $5.00. Plenty of parking, food. Check in on 146.52 or 147.90/30. For information, tickets or tables: Paul Kilzer, W8GAX, 393 Pole Lane Road, Marion, Ohio 43302.

NEW MEXICO: The UNM ARC and Westside ARC are cosponsoring a tailgate swapfest, November 3, 10 AM to 2 PM MST, UNM North Campus parking lot, corner of University Blvd. and Tucker Avenue, Albuquerque. No charge. Bring own tables. Talk in on 147.75/147.15 and 449.3/444.3 repeaters. For further information SASE to K8BI, WB5YYX or WA5WHN or via 3.939 MHz, 0100 UTC daily.

MICHIGAN: The Blossomland Amateur Radio Association's 1984 Blossomland Blast, Sunday, October 7, Lake Michigan College Community Center, I-94 exit 30, west of Benton Harbor. 8 AM to 3 PM EDT. Admission $3 per person. Special features: Air Force MARS display, Skywarn training program, RC airplane display. Talk in 22/82 and 52. For table space: BARA, POB 175, St. Joseph, MI 49085 or Paul, WD8MWT (616) 983-1710.

ILLINOIS: Third annual CCRL Hamfest, Sunday, October 21, 7 AM to 2 PM, American Legion Post #21, 6040 N. Clark St., Chicago. Admission $1.00 advance, $1.50 at door. $2.00 per table. Talk in on 145.030 simplex. For information: Norman Geuder, KA9EZA, John Ibes, KA9FUI or Frank Bonnell, WB9OHN.

MASSACHUSETTS: The Framingham Amateur Radio Association's annual Fall flea market, Sunday, October 28, Framingham Civic League Bldg., 214 Concord Street, Framingham. Doors open 10 AM. Sellers setup 8:30. Admission $2.00. Tables $10.00 pre-registration only. Bargains galore. Talk in on 75/15 and 52. Contact: Jon Weiner, K1VVC, 52 Overlook Drive, Framingham, MA 01701. (617) 877-7166.

NEW ENGLAND: Hosstraders' Fall Tailgate Swapfest, Saturday, October 6, sunrise to sunset at Deerfield, NH Fairgrounds. Admission $2 including tailgaters. Friday night camping at nominal fee after 4 PM. No reservations. Profits benefit Boston Burns Unit of Shriners Hospital. Last Spring's donation $5,813.00. For map to northeast's biggest ham flea market SASE to Norm, WA1IVB, RFD Box 57, West Baldwin, ME 04091.

OPERATING EVENTS

"Things to do..."

OCTOBER 6-14: Special event stations K5MHZ and KN5D will operate during the 13th annual International Hot Air Balloon Fiesta in Albuquerque, New Mexico. Most operations SSB with some RTTY, CW and SSTV. Variable hours. Frequencies: 3900, 7230, 14250, 21350 and 28550 and 147.510 simplex FM. For a special QSL send QSL to PO Box 997, Corrales, NM 87048. Include SASE with sufficient postage.

OCTOBER 6: Southeastern Michigan ARA will sponsor Michigan All Saints Day, 1500Z to 2100Z with stations in all 9 cities named after Saints. Phone only, General portions of 20, 40 and 60 meters plus 147.480 and/or nearest repeater. Look for St. Joseph, KC8JX, St. Helen N8BAR, St. Johns NI6L, St. Clair W8GV, St. Charles WB8TTA, St. Ignace KD8CW, St. Louis WA8AEG, St. James KD8CG and Sault Sainte Marie WA8DLO. QSL once with #10 SASE and log to N8COY, 161 Lothrop, Grosse Pointe, MI 48236 for certificate with endorsements.

OCTOBER 16, 17, 18: Colquitt County Ham Radio Society will operate club station WD4KOW from the site of the 7th annual Sunbelt Agricultural Exposition. 0900 to 1700 EDST each day. Operations: General portion of HF bands. Members will listen for visiting hams on 146.19/.79. For a special QSL card send SASE to: CC Ham Radio Society, PO Box 813, Moultrie, GA 31776.

OCTOBER 27 AND 28: Members of the Rutherford Appleton Laboratory ARC (G3RRS) will be active from VP2MF during this years contest. Operators: G3SJK, G3UKS, GM3YOR, G4BGH, G4JVG, G4XRI, G4XRJ. An award for DX stations working VP2MF on 10-160m during the contest will be issued. QSL via bureau or to G3RRS, c/o Jean Mills, R20, Rutherford Appleton Laboratory ARC, Chilton, Didcot, Oxon, UK.

OCTOBER 20 AND 21: The Armadillo Gang will operate WD5HOR to commemorate the Arkansas Pass "Shrimporee". Operation will be on 10, 15, 20, 40, 80 meters 15kc from lower General band edges. QSL card available via SASE to the Armadillo Gang, WB5YPE, David Stephens, 5709 Bobalo, Corpus Christi, TX 78412.

NOVEMBER 25 AND 26: The BOMB Squad (Best of Mt. Baldy) will operate W6HCP (Hollywood Christmas Parade) from 1800Z, November 25 to 0400Z, November 26. Frequencies: 7.284, 14.284, and 21.284 MHz SSB. SASE to W6GVR for special commemorative QSL.

OCTOBER 14: LARC will operate special event station WA3OGA from 1300Z to 2400Z to commemorate the Bollman Truss RR Bridge — the only one of its kind in the world. Frequencies: 7237, 14285, 21385, 144.250 USB and 147.540 simplex. Certificate available for #10 SASE to LARC, PO Box 3039, Laurel, MD 20708.

OCTOBER 21, 22: JOTA — Scouts 27th annual Jamboree on the Air. Look for K2BSA, the BSA headquarters station in Dallas, Texas and HB9S, the World Scout Headquarters in Switzerland and other special call signs from many countries. Calling frequencies: CW — 3590, 7030, 14070, 21140, 28190. Voice — 3940, 7290, 14290, 21360, 28990. RTTY, SSTV, ATV on usual frequencies. Check Novice frequencies. Do a good turn for Scouting and Ham Radio.

NOVEMBER 3: K4MJN will operate a special events station in Blythewood, SC, to commemorate the birthplace of J. Gordon Coogler, acclaimed by literary critics as the "worst practicing poet in U.S. literary history!". All stations working K4MJN during this second annual festival will receive a handsome certificate with a photo of "The Bard of Blythewood" and some of his poetry. Please send QSL and contact number with large SASE to K4MJN, Rt. 3, Box 154, Blythewood, SC 29016. 14.290 MHz from 1400Z to 1800Z and 21.390 MHz from 1800Z to 2200Z.

OCTOBER 20: The 24th Infantry Division Association will sponsor a special event station K4TF, to commemorate the 40th anniversary of the landing in the Philippines. A special commemorative certificate to any Amateur station making 2-way contact with K4TF during the 24-hour GMT period. Operations will be 10 kHz inside the General portion of each band. For certificate send QSL card and 9 x 12 SASE to K4TF, 1630 Venus Street, Merritt Island, Florida 32953.

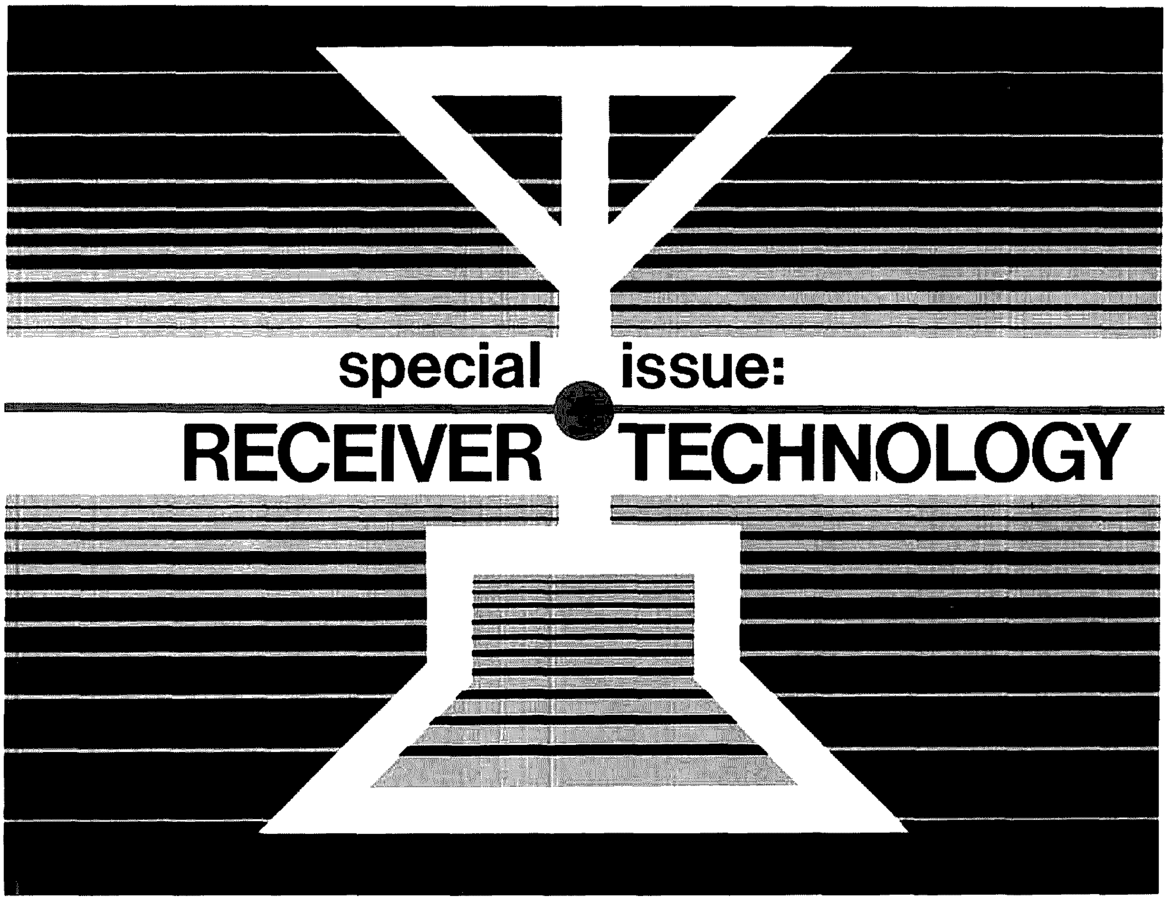
AMSAT Technical Symposium and General Membership Meeting, Sunday, November 10, Amfac Hotel, 8601 Lincoln Blvd., Los Angeles, CA 90045. Tech presentations on present/future Amateur satellite projects. General membership meeting following a banquet dinner. For conference registration information SASE to Dennis Dinga, N6DD, PO Box 4111, Diamond Bar, CA 91765.

OCTOBER 13 AND 14: Oregon QSO party sponsored by the Hermiston ARC, 1700Z Oct. 13 to 0800Z and from 1500Z Oct. 14 to 0000Z Oct. 15. Exchange: OR stations, signal report and county. Others signal report and state/province/country. Mixed mode or CW only. You may obtain log sheets from HARC (please SASE). Log sheets must be received by November 12. Mail entries and request log sheets from: Hermiston ARC, PO Box 962, Hermiston, OR 97838.

OCTOBER 13 AND 14: QRP Amateur Radio Club International Fall QSO Contest. 1200 UTC Oct. 13 to 2400 UTC Oct. 14. 24 hours max. operation. Only one mode of operation, CW or SSB may be used. Exchanges: Members give RS(T), state/province/country and QRP ARCI membership number. Non-members give RS(T), state/province/country and power output. Logs must be received by November 12. Send all material to: QRP ARCI Contest Chairman, Gene Smith, KA5NLY, 8201 Chatham Drive, Little Rock, AR 72207.

NOVEMBER 1984 / $2.50

ham radio magazine



hr

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ham radio magazine

NOVEMBER 1984

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contents

REFLECTIONS

polish until it shines

It was a lazy afternoon. The air was warm, the sky was blue, and a soft sea breeze wafted gently across the deck. The place was Martinique, and a young radio operator from the SS *Brasil* — me — had the afternoon free. In those days my call sign was WMDT (all ships used four letters for identification), and I was at the halfway point in my fourth trip out to sea as a radio operator in the U.S. Merchant Marines.

Thinking back now, I recall the slow, undulating motion of the ship, the immense expanse of ocean, and the fresh smell of sea breeze created by the water splashing against the fantail. It was a wonderful experience for a lad of 19 to be able to visit many foreign ports, operate a high-power shipboard radio station (with four receivers), to receive room and board — and *be paid* — for the privilege!

On that lazy afternoon I decided to visit my counterparts (radio operators) aboard the SS *France* (FNRR). I suppose it was natural to want to see what equipment and antennas they had, what operating procedures they used, and in general, what their life was like aboard ship.

While the radio room on the *France* was larger than the *Brasil's*, they had about the same complement of transmitters and receivers as we had aboard our vessel, plus a high-resolution TV system used to pipe signals throughout the ship. Although our working conditions seemed similar, our feelings seemed to be quite different. The radio operators (there were about six, I believe) all appeared to be good, close friends, and they obviously enjoyed each other's company. I couldn't help but compare the atmosphere aboard my ship with that of the *France*. Though we were all friendly while on duty, we went our separate ways immediately after docking — I guess you could call our style "rugged individualism." I found myself preferring, however, the camaraderie shown by my new-found friends aboard this "foreign" liner.

What is a visit to France (or a French ship) without tasting the food? I was invited to lunch. In the cafeteria we enjoyed an excellent meal, several glasses of good wine, and amicable conversation. But suddenly my attention focused on one of the kitchen workers. I couldn't help noticing the considerable effort he was applying to the polishing of his equipment. Summoning up my best French, I went over to him and asked why he worked so hard. Were they that strict aboard the ship?

First he laughed. Then he became quite serious and said something that I'll probably never forget: "This is my job. I want to do the best I can at it. If I thought it were 'beneath me' to do this job, I'd get another."

I couldn't help thinking how many people I knew and had known who had what might be considered very good jobs, yet complained, for one reason or another, that they should have been doing something else. We have so much in this wonderful country of ours. We have resources and resourcefullness. Our children have the opportunity to acquire an excellent education, and we have the facilities to train them — and ourselves — for many different interesting jobs.

In Amateur Radio it's no different. We have the equipment, spectrum, technical resources, and obviously the time (just listen to some of our lengthy rag chews!) and yet I often come away from an evening on the air with the feeling that something's missing. We're all, it appears, "rugged individuals" diligently protective of our own frequencies and thoughts, content to do the same thing day after day. (For those who know my operating habits, perhaps I'm a fine one to talk . . . I *do* zero in on chasing quite a bit of DX.) I guess what I'm trying to say is that I'd be very happy to see what we have appreciated more and used more fully.

For my part I'm going to continue my experiments in antenna development and propagation studies, my two favorite technical subjects. But first I'm going to work on a more pressing problem — how to squeeze just two more hours into a 24-hour day. I don't think that's asking for too much.

Rich Rosen, K2RR
Editor-in-Chief

presstop de W9JUV

REALLOCATION OF THE TOP HALF OF 160 METERS TO RADIOLOCATION could take place in the very near future. In a mid-September Notice of Proposed Rule Making, the Commission has proposed moving non-government radiolocation operations from their present slot between the top end of AM broadcast and the bottom of 160 up to 1900-2000 kHz. The shift is based on the WARC '79 upward expansion of AM broadcast, displacing present radiolocation operation.

Ironically, The Importance Of Medium Frequency Radiolocation is being questioned in a Petition for Initiation of Inquiry Procedure filed by the ARRL just the day before the FCC's NPRM was released. In it the League asks that the actual spectrum requirements of the individual radiolocation users be specified along with the actual number of such stations that might be active in any geographical area. Though the ARRL petition addressed the needs of all non-government radiolocation, it specifically asked the Commission to consider whether radiolocation's real needs are sufficient to justify taking over the 1900-2000 kHz slot.

The League Has Now Petitioned The FCC To Withhold Consideration of the reallocation docket until after it considers the League's Inquiry Procedure petition.

A BILL STRENGTHENING FEDERAL LAW ON MALICIOUS INTERFERENCE has been introduced in the U.S. Senate by Barry Goldwater, K7UGA. In his bill, S-2975, Sen. Goldwater would make any operator of equipment used to maliciously interfere with any form of radio communications (or radar) subject to Section 501 of the Communications Act if he continues after receipt of written notice to stop. Section 501 provides for fines up to $10,000 and two years in prison; under present law the fine for malicious interference is only $500. In addition, the equipment used to generate the malicious interference could also be siezed.

RELIEF OF AMATEUR OPERATIONS FROM STATE AND LOCAL REGULATION is being sought by the ARRL. The League has asked the FCC to issue a "Declaratory Ruling of Limited Federal Preemption of State and Local Regulation of Amateur Radio Station Installation and Operation," to spell out just what limitations local and state authorities could place over federally-licensed Amateurs. A similar request regarding local regulation of TVRO satellite dishes was filed some time ago by United States Communications, Inc.

Comments From Concerned Amateurs, Particularly Those who've had problems with local regulators, are being sought by the Commission. An original and four copies should go to the Secretary, FCC, 1919 M St., NW, Washington, D.C. by November 9; refer to PRB-1. A copy of those Comments, along with any supporting documentation, would also be very helpful to the ARRL in its efforts. USCI's proposal on behalf of TVRO owners has generated strong opposition from a number of governmental organizations, and it's almost certain they'll resist the League's request with equal fervor.

THOUGH THERE'S BEEN NO REAL CHANGE IN THE 220 MHZ SITUATION since last month's Presstop, there have been some interesting developments. "220 Notes" Publisher K9XI has requested a Congressional investigation of the FCC's Office of Science and Technology, based on concerns that the OST may have been improperly involved in the STI petition that asked for reallocation of the 220-222 MHz slot to ACSB. "Westlink" reports Congress is getting plenty of mail on the subject, with Sen. Goldwater's office receiving about a thousand letters from concerned Amateurs and California Senator Pete Wilson almost 400.

WA2MCT's Petition To Permit Novices All-Mode 220 Privileges has been denied and dismissed by the FCC. In denying the petition Private Radio Bureau Chief Bob Foosaner noted that both the FCC and National Telecommunications Information Administration (NTIA) are conducting on-going studies of future 216-225 MHz uses, so it is "not appropriate to consider petitions which could have a major impact on the 220 MHz band..." at this time.

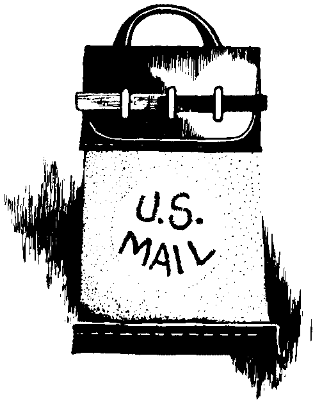
A SPREAD SPECTRUM FREQUENCY HOPPING 2-METER BEACON IS NOW ON THE AIR from Falls Church, Virginia. Start and stop frequencies are 144.5 and 147.7 MHz, on a 25-kHz spaced pseudo-random pattern. It's transmitting MCW on narrow band FM with a hop rate of 10 hops per second, sending a series of Vs followed by the station ID. Contact N4EZV for details.

EXTENSIVE CHANGES IN THE VEC PROGRAM HAVE BEEN PROPOSED by W6NLG on behalf of the Sunnyvale VEC Amateur Radio Club, the newly appointed California VEC. They'd like the prior notification requirement relaxed, and more leniency with respect to the exams Advanced class VE's can administer. They'd also limit any VEC to a maximum of 3 call areas, to provide for local control. An RM number has not been assigned at the present time.

It Appears The FCC May Let The VEC Program Run As Is for the time being, until both it and the participants have enough experience to know what (if any) real bugs it has. However, it may act favorably on RM-4835, which would shorten the delay period for retaking a failed exam from the present 30 days to 7, despite ARRL opposition.

ARIZONA IS ADOPTING 20 KHZ SPACING ON 2 METER'S TOP END, effective immediately. No more "odd digit" coordination for either new repeaters or for changes in existing machines will be permitted, and a statewide program to move all odd digit systems will begin soon.

AN NPRM TO IMPLEMENT VARIOUS WARC BANDS IS DUE for FCC release very soon, possibly before this sees print. It's expected to include 24 and 902 MHz as well as 10 MHz (still operating under temporary authorization), and probably other WARC changes as well.



comments

wait for the mailman

Dear HR:

Thanks for bringing VHF and UHF out of the dark ages and into the daylight. As I sat here carefully cutting out W1JR's article on propagation, (July, 1984) it occurred to me that it's the best primer I've ever read. The article now has a home on my research book shelf right next to authorities such as *Natural Electromagnetic Phenomena*, *Electronic Density Profiles in the Ionosphere and Exosphere*, and other noteworthy journals and papers.

WB3BGU's series on VHF and UHF Antenna Design (May-October, 1984) helps clear the smoke screen on design that has snowed hams for years. I have built and put up some large arrays over the years; Stan's notes are the best guide ever written for hams.

Since Rich Rosen took over as Editor-in-Chief, *ham radio* has moved to the number 1 position on my wait-for-the-mailman list. Keep up the great work.

Sid Liberman, WA2FXB
Woodbridge, New Jersey

Model 28 printer

Dear HR:

I have a TRS80 Color Computer,™ Kantronic Software, and an MFJ TU-1224. I'd like to use my Model 28 as a printer. Can someone out there show me how?

John L. Gill
6000 Duda Road
House Springs, Missouri 63051

cheers

Dear HR:

Regarding your July, 1984 editorial, "The Number 1 Question," thanks! Not exactly for spelling out how to write a magazine article, but for announcing the birth of the "Superduper Louden-Boomer Metal Noodle." We in this area are using the "new and improved" version with extraordinary results and will shortly — yesterday, I believe — come out with an even more versatile one —the DASH 2 — on which I would be glad NOT to write a technical paper.

Seriously, though, I enjoy *ham radio* very much. Keep up the good work!

Frank Brumett, WB4CIZ
Lexington, Kentucky

wideband VCO design

Dear HR:

Your July, 1984, issue came just in time. I was showing my students how to use the Smith chart for finding the length of a transmission line to act as an inductor and I wanted a circuit to build. Alan Victor's article on wideband VCO design was just what I needed.

The circuit was easy to build, and because the resonator is shielded, it was immune to handling by the students. I was able to vary the frequency of the Colpitts oscillator throughout the FM radio band for the students to hear. This circuit helped my students in applying theory to a practical application.

Joe Avampato, W8DKR
Fort Mill, South Carolina

In the May, 1984, article, "Remote-controlled 40, 80, and 160-meter Vertical," reference was made to 4-inch O.D. irrigation pipe. Local inquiries produced the following information: 4-inch aluminum irrigation pipe with 0.050 inch wall is available in lengths up to 40 feet from Larchmont Engineering, P.O. Box 66, 11 Larchmont Lane, Lexington, Massachusetts 02173. The price is $2.38 per foot; other sizes are available. (Check your local phone book for additional sources.)

For additional sources of Ledex, also specified in W7LR's article, send an SASE to *ham radio*, Greenville, New Hampshire 03048.

Editor

quiet! preamp at work

Understanding preamplifiers means understanding all the important parameters of receiver performance

For years, the **standard technique** employed by Radio Amateurs to improve receiver sensitivity has been to precede their receivers with one or more stages of preamplification. Invariably a preamplifier that performs well on the bench will actually *degrade* the actual on-the-air system sensitivity. This article explores the relationship between gain, noise figure, bandwidth, distortion, and sensitivity in an attempt to answer the classic preamp question, "If a little is good, is a lot better?"

sensitivity

Sensitivity is a measure of the weakest input signal that will produce a specified output signal-to-noise ratio. We can quantify receiver performance in terms of *minimum discernible signal* sensitivity, which is the input level producing an output signal-to-noise ratio of unity; *tangential signal sensitivity* , which generally refers to the input level needed to produce an output signal-plus-noise to noise ratio of 6 dB or the RF level required to produce a detected signal which is 8 dB above the RMS noise level[1]; or *threshold*, which refers to the input amplitude required to produce a specified level of receiver quieting and is frequently employed in FM systems. All of these sensitivity measures are a function of the receiver circuitry's internally generated noise, bandwidth, and distortion.

Of these three parameters, the receive bandwidth can be considered fixed for a given application, and would ideally be wide enough to pass all the modulation sidebands of the desired signal, yet sufficiently narrow to exclude both background noise and any adjacent-channel signals. Because the response bandwidth of modern receivers is established primarily in the IF stages, it is relatively independent of the parameters of any preamplifier employed.

Both noise and distortion, on the other hand, are very much influenced by preamplifier performance. Most Radio Amateurs are now aware that preamplifier gain, by itself, does not necessarily assure an improvement in receiver sensitivity. Rather, to be beneficial in a system, the preamplifier must generate an internal noise level significantly lower than that generated by the receiver it precedes. The noise relationships in a cascade of stages are quantified by the now-familiar Friis Equation.[2] A well-known rule of thumb derived from the Friis Equation is that if a preamp's gain exceeds by at least 10 dB the noise figure of the receiver it precedes, the noise performance of the preamplifier will dominate the cascade.

Yet the above relationship serves merely to confuse the Amateur who measures a new preamp at a regional VHF Conference at, say, 3 dB noise figure for 15 dB gain, brings it home, installs it in front of a 10 dB noise-figure receiver, and finds its sensitivity actually degraded. What has been overlooked? Probably the effects of distortion.

distortion

A linear amplifier is one whose output signal is an exact replica of the input signal, measured in either

By **H. Paul Shuch, N6TX**, 14908 Sandy Lane, San Jose, California 95124

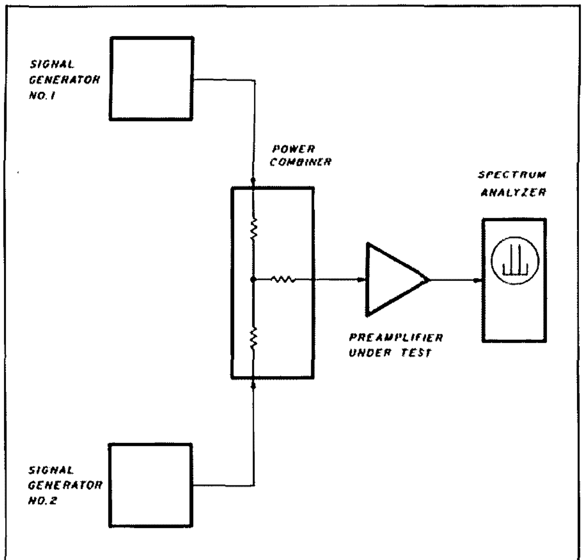


fig. 1. Two-tone intermodulation distortion test configuration.

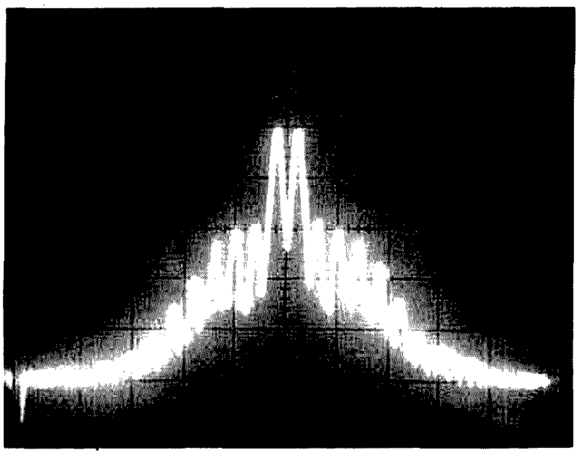


fig. 2. Typical intermodulation distortion spectrum display. Note the next pair of "signals" (IMD) are 20 dB down from the primary two-tone output.

the time or frequency domains, differing only in its increased amplitude. Try as we might, we cannot build truly linear amplifiers in the real world. Any non-linearity introduced by an amplifier will manifest itself as a deviation from sinusoidal response when viewed in the time domain, or as the generation of new frequencies when measured in the frequency domain.

In a receive preamplifier, as in any non-linear device, the distortion products generated are integer multiples (harmonics) of the input frequency, plus their various sums and differences. Normally these distortion products would not degrade receiver sensitivity, as they would fall outside of the receiver's passband. Rare, however, is the receiver to which only a single input signal is applied. In our crowded spectra, we can anticipate countless signals of varying amplitudes within the passbands of our preamplifiers, only one of which (at a time) can be said to constitute "signal." All potentially interfering waveforms must, from a communications standpoint, be classified as noise.

It is these multiple input signals that give rise to both intermodulation (mixing of in-band signals) and cross-modulation (mixing of signals from in-band with out-of-band) distortion. When the harmonics of one signal mix with the harmonics of another, the resulting distortion products can fall within the receiver passband, degrading sensitivity.

dynamic range

Neglecting distortion effects, the weakest signal to which a receiver can respond is a function of its bandwidth and noise performance. If the multiple input signals applied to a receive system are all relatively low in amplitude, their distortion products may fall below this sensitivity limit, and be negligible. But if the input signals are of sufficient amplitude, their distortion products may appear strong enough to degrade reception of the desired signal. Thus, noise figure of a receiver generally determines the weakest signal to which it can respond. Maximum spurious free input signal, a function of a receiver's linearity, establishes an upper limit for the range of signal amplitudes to which the receiver can respond without generating perceptible distortion. The difference between sensitivity and maximum spur-free input levels is called spurious-free dynamic range, and represents a primary limitation in receiver performance.

Dynamic range is generally degraded by the addition of a preamplifier in front of a receiver. Although the low inherent circuit noise of a preamplifier may significantly improve minimum discernible signal sensitivity, degradation occurs because any additional gain in a system increases the amplitude of the desired signal, but increases the amplitude of the distortion products at an even greater rate, diminishing the maximum spurious-free input signal level. Thus, at least with respect to preamplifier gain, the old axiom, "If a little is good, a lot is better" can get us into trouble. Preamplifiers should be used only when actually necessary to improve weak-signal performance, and then only with as much gain as is actually necessary to establish the required system noise performance.

Even so, preamplifiers can result in a net degradation in system sensitivity. Some preamps are worse than others in this respect; as far as dynamic range is concerned, not all preamps are created equal. We need to measure and quantify their dynamic range, as well as their noise figure, in order to accurately predict their impact on system performance.

```
10 '---------> IMD.BAS <-------------------------
20 '             Rev. A, 24 May '84
30 '             by N6TX
40 '  COPYRIGHT (C) 1984 MICROCOMM
50 '
60 '  Determines Spurious-Free Dynamic Range from
70 '  Spectrum Analyzer Two-Tone IMD Measurements
80 '
90 '-----------------------------------------
100 CLR$ = CHR$(26) ' Defines Clear-Screen String
110 PRINT CLR$
120 PRINT "DO YOU WISH OUTPUT ROUTED TO:"
130 PRINT
140 PRINT "                        PRINTER (P)"
150 INPUT "                 or     SCREEN  (S)";PR$
160 IF PR$="P" OR PR$="p" OR PR$="S" OR PR$="s" GOTO 200
170 PRINT CLR$
180 PRINT "YOU MUST RESPOND WITH 'P' OR 'S'" : PRINT
190 GOTO 120
200 '-----------------------------------------
210 PRINT CLR$
220 PRINT "TONES -->            |   |"
230 PRINT "                     |   |"
240 PRINT "                     |   |"
250 PRINT "                     |   |"
260 PRINT " IMD  -->        |   |   |   |"
270 PRINT "                 |   |   |   |"
280 PRINT "             ----------------" : PRINT
290 PRINT "FOR TWO-TONE OUTPUT SPECTRUM AS INDICATED ABOVE,"
300 PRINT
310 INPUT "ENTER Two-Tone Output Amplitude, in dBm ",TONES
320 INPUT "ENTER Third-Order IMD Amplitude, in dBm ",IMD
330 INPUT "ENTER System Gain,               in dB  ",GAIN
340 INPUT "ENTER System Noise Figure,       in dB  ",NF
350 INPUT "ENTER System Bandwidth,          in kHz ",BW
360 I3=TONES + ((TONES - IMD) / 2)
370 MDS = NF - 144 + 10*(LOG(BW)/LOG(10))
380 INMAX = ((2/3)*(I3-GAIN))+(MDS/3)
390 RANGE = INMAX - MDS
400 '-----------------------------------------
410 PRINT CLR$
420 PRINT "INTERMODULATION ANALYSIS BY MICROCOMM"
430 PRINT : PRINT
440 PRINT USING "####.# dBm -->          |   |    ";TONES
450 PRINT       "                        |   |    "
460 PRINT       "                        |   |    "
470 PRINT       "                        |   |    "
480 PRINT USING "####.# dBm -->      |   |   |   |";IMD
490 PRINT       "                    |   |   |   |"
500 PRINT       "                 ----------------"
510 PRINT : PRINT
520 PRINT USING "SYSTEM GAIN          = ####.# dB";GAIN
530 PRINT USING "SYSTEM NOISE FIGURE  = ####.# dB";NF
540 PRINT USING "SYSTEM BANDWIDTH     = ####.# kHz";BW
550 PRINT
560 PRINT USING "OUTPUT THIRD ORDER INTERCEPT POINT = ####.# dBm";I3
570 PRINT USING "MINIMUM DISCERNIBLE INPUT SIGNAL   = ####.# dBm";MDS
580 PRINT USING "MAXIMUM SPURIOUS-FREE INPUT SIGNAL = ####.# dBm";INMAX
590 PRINT USING "SPURIOUS-FREE DYNAMIC RANGE        = ####.# dB";RANGE
600 '-----------------------------------------
610 IF PR$="S" OR PR$="s" THEN GOTO 800
620 LPRINT "INTERMODULATION ANALYSIS BY MICROCOMM"
630 LPRINT : LPRINT
640 LPRINT USING "####.# dBm -->         |   |    "; TONES
650 LPRINT       "                       |   |    "
660 LPRINT       "                       |   |    "
670 LPRINT       "                       |   |    "
680 LPRINT USING "####.# dBm -->     |   |   |   |"; IMD
690 LPRINT       "                   |   |   |   |"
700 LPRINT       "                ----------------"
710 LPRINT : LPRINT
720 LPRINT USING "SYSTEM GAIN          = ####.# dB";GAIN
730 LPRINT USING "SYSTEM NOISE FIGURE  = ####.# dB";NF
740 LPRINT USING "SYSTEM BANDWIDTH     = ####.# kHz";BW
750 LPRINT
760 LPRINT USING "     OUTPUT THIRD ORDER INTERCEPT POINT = ####.# dBm"; I3
770 LPRINT USING "     MINIMUM DISCERNIBLE INPUT SIGNAL   = ####.# dBm";MDS
780 LPRINT USING "     MAXIMUM SPURIOUS-FREE INPUT SIGNAL = ####.# dBm";INMAX
790 LPRINT USING "     SPURIOUS-FREE DYNAMIC RANGE        = ####.# dB";RANGE
792 LPRINT
794 LPRINT
796 LPRINT
798 LPRINT
800 '-----------------------------------------
810 PRINT : PRINT
820 INPUT "TYPE <return> TO CONTINUE, 'Q' TO QUIT ",O$
830 IF D$ = "Q" OR D$ = "q" THEN GOTO 850
840 GOTO 100
850 '
860 '---------------------------------------------------------------
870 '
880 '          EQUATIONS EXECUTED
890 '
900 'MINIMUM DISCERNIBLE SIGNAL  =  -174 dBm/Hz  + NF (dB)  + 10 * LOG BW (Hz)
910 '
920 'OUTPUT INTERCEPT POINT      =  P (tones)  +  [ P (tones) - P (imd) ] /  2
930 '
940 'MAXIMUM INPUT SIGNAL LEVEL  =  (2/3) * (INTERCEPT - GAIN) + (M. D. S. /3)
950 '
960 'SPURIOUS FREE DYNAMIC RANGE =  MAXIMUM INPUT - MINIMUM DISCERNIBLE SIGNAL
970 '
980 '
990 '---------------------------------------------------------------
1000 END
```

fig. 3. CP/M BASIC language program listing to determine spurious-free dynamic range from spectrum analyzer two-tone IMD measurements.

gain compression

Inferences about an amplifier's dynamic range can be drawn by applying to its input a single signal of varying amplitude and observing the amplitude present at the output. In its linear region, the amplifier will produce a 1-dB change in output signal amplitude for every 1-dB change in the applied signal. That is, the gain of the amplifier is independent of applied signal level. But as the upper limit of dynamic range is approached, output signal changes will be unable to keep pace with the input. That is, the gain of the amplifier compresses at the upper end of its dynamic range. The output level at which the amplifier is exhibiting 1 dB less gain that it was under weak-signal conditions is referred to as its *output 1-dB compression point*, and is an indicator of the amplifier's immunity to intermodulation and cross-modulation distortion.

```
INTERMODULATION ANALYSIS BY MICROCOMM

  -6.0 dBm -->          |   |
                        |   |
                        |   |
                        |   |
 -26.0 dBm -->      |   |   |   |
                    |   |   |   |
                 ----------------

SYSTEM GAIN          =   -6.0 dB
SYSTEM NOISE FIGURE  =    7.0 dB
SYSTEM BANDWIDTH     =    2.4 kHz

     OUTPUT THIRD ORDER INTERCEPT POINT =    4.0 dBm
     MINIMUM DISCERNIBLE INPUT SIGNAL   = -133.2 dBm
     MAXIMUM SPURIOUS-FREE INPUT SIGNAL =  -37.7 dBm
     SPURIOUS-FREE DYNAMIC RANGE        =   95.5 dB
```

fig. 4. IMD analysis of a double-balanced mixer with a +7 dBm injected LO level.

For a given noise figure, the preamplifier with the highest compression point will offer the greatest spurious-free dynamic range. But correlating the two parameters directly is difficult because the relationship between compression and distortion varies between active devices, and between circuit configurations.

Another indicator of dynamic range relates to the fact that if you continue to increase the drive level to an amplifier beyond the compression point, the gain further decreases. Eventually, the amplification of the desired signal is degraded to a point at which its amplitude at the output of the amplifier, and those of the intermodulation distortion products, would be the same. The output level at which this should occur is called the *output intercept point.** Intercept point is more readily correlated to dynamic range than is compression point, but is difficult to measure directly. To best quantify dynamic range limitations, it is necessary to test the preamplifier in its actual operating environment — that is, under multiple-signal conditions.

two-tone testing

In the method of dynamic range testing prevalent in industry, two sinusoidal signals of equal amplitude are applied to the input of the device under test, and the resulting output spectrum monitored in the frequency domain. The two input signals, or tones, may be generated by summing the outputs of the two signal generators in a power combiner, or by applying a single RF source to the LO input of a balanced mix-

*For many amplifier circuits, this is a *theoretical*, rather than an *attainable*, level, because the active device may burn out before this output level is reached.

```
INTERMODULATION ANALYSIS BY MICROCOMM

 -7.0 dBm -->           | |
                        | |
                        | |
                        | |
 -65.0 dBm -->        | | | |
                      | | | |
                    ----------------


SYSTEM GAIN            =   -7.0 dB
SYSTEM NOISE FIGURE    =    8.0 dB
SYSTEM BANDWIDTH       =    2.4 kHz

    OUTPUT THIRD ORDER INTERCEPT POINT =    22.0 dBm
    MINIMUM DISCERNIBLE INPUT SIGNAL   =  -132.2 dBm
    MAXIMUM SPURIOUS-FREE INPUT SIGNAL =   -24.7 dBm
    SPURIOUS-FREE DYNAMIC RANGE        =   107.5 dB
```

fig. 5. IMD analysis of a double-balanced mixer with a +17 dBm injected LO level.

```
INTERMODULATION ANALYSIS BY MICROCOMM

 -14.0 dBm -->          | |
                        | |
                        | |
                        | |
 -25.0 dBm -->        | | | |
                      | | | |
                    ----------------


SYSTEM GAIN            =   12.0 dB
SYSTEM NOISE FIGURE    =    2.0 dB
SYSTEM BANDWIDTH       =    2.4 kHz

    OUTPUT THIRD ORDER INTERCEPT POINT =    -8.5 dBm
    MINIMUM DISCERNIBLE INPUT SIGNAL   =  -138.2 dBm
    MAXIMUM SPURIOUS-FREE INPUT SIGNAL =   -59.7 dBm
    SPURIOUS-FREE DYNAMIC RANGE        =    78.5 dB
```

fig. 6. IMD analysis of a bipolar junction transistor preamplifier.

er, a suitable audio signal generator to the mixer's IF input, and applying to the device under test the double-sideband (two-tone) signal appearing at the mixer's RF port. In either case, the two tones must be separated in frequency sufficiently to be individually resolved on the spectrum analyzer's display, yet sufficiently close in frequency to both fall within the response bandwidth of the device under test.

A typical interconnection of instruments for two-tone dynamic range analysis is shown in **fig. 1**, and a typical resulting spectrum is displayed in **fig. 2**. Note that the distortion products of greatest amplitude (in this case, the pair of signals immediately adjacent to the two applied tones) are roughly two divisions, or 20 dB, below the amplitude of the desired output tones. The intermodulation distortion level of this particular amplifier, measured at this particular signal level, is thus -20 dB.

If the vertical axis of the spectrum analyzer is calibrated in absolute amplitude (typically in dBm), the output power per tone, the PEP output power (6 dB above the level of each individual tone), and power of the individual distortion products can be readily determined. And from these values, with minimal number crunching, we can determine the dynamic range of the preamplifier.

data analysis

The mathematical relationships applied next are, as is said in college texts, "beyond the scope of this course." However, I have included in **fig. 3** a listing of a Micro-soft™ BASIC program that performs the complete analysis. Although written to run under the CP/M™ operating system, the program can likely be modified to run on any of the popular home computers using their version of BASIC. **Figures 4** through **8** are sample executions of the IMD program for various receiver configurations. Comparing these printouts will enable us to draw some significant conclusions with regard to the utility of preamplifiers in VHF and UHF communications systems.

mixer design considerations

As a rule, balanced mixers offer excellent dynamic range and intermodulation distortion performance, although their weak-signal sensitivity leaves something to be desired. Mixers are designed to operate at different levels of local oscillator injection, and generally, the higher the LO level employed, the higher will be the mixer's compression level. However, raising the LO injection above perhaps 5 milliwatts tends to degrade mixer conversion efficiency and noise figure. Nonetheless, as **figs. 3** and **4** indicate, so-called high level mixers offer sufficiently improved dynamic range to override the considerations of slightly degraded sensitivity, in most applications.

Not shown in the computer runs, but worthy of consideration, are the so-called "starved LO" mixers. These devices use an extremely low LO injection level with external DC bias of their mixer diodes, and excel in low-noise performance. Their dynamic range, however, is severely degraded, typically 12 to 15 dB below that of even the "low-level" balanced mixer shown in **fig. 3**. Thus, except in those applications in which it is impractical to generate 5 milliwatts or more of LO injection, starved LO operation should be avoided.

The same is true for harmonic mixers. These devices are extremely popular in microwave TV receive converters, and employ LO injection at half the normal frequency, with the mixer diodes serving double duty as frequency multipliers. Obviously, the more frequencies we generate within a mixer, the more spurs will be available to bite us later. I recommend multiplying in a stage separate from that doing the heterodyne conversion.

preamp design considerations

Most receive preamplifiers operate with their active devices drawing relatively low quiescent current. This is done because high device current generates high thermal activity, which degrades noise performance significantly. Unfortunately, biasing any active device

```
INTERMODULATION ANALYSIS BY MICROCOMM

 -9.0 dBm -->            | |
                         | |
                         | |
                         | |
 -22.0 dBm -->         | | | |
                       | | | |
                  -----------------

SYSTEM GAIN          =    17.0 dB
SYSTEM NOISE FIGURE  =     3.0 dB
SYSTEM BANDWIDTH     =     2.4 kHz

     OUTPUT THIRD ORDER INTERCEPT POINT =    -2.5 dBm
     MINIMUM DISCERNIBLE INPUT SIGNAL   =  -137.2 dBm
     MAXIMUM SPURIOUS-FREE INPUT SIGNAL =   -58.7 dBm
     SPURIOUS-FREE DYNAMIC RANGE        =    78.5 dB
```

fig. 7. IMD analysis of a MOSFET preamplifier.

```
INTERMODULATION ANALYSIS BY MICROCOMM

 -2.0 dBm -->            | |
                         | |
                         | |
                         | |
 -28.0 dBm -->         | | | |
                       | | | |
                  -----------------

SYSTEM GAIN          =    24.0 dB
SYSTEM NOISE FIGURE  =     1.0 dB
SYSTEM BANDWIDTH     =     2.4 kHz

     OUTPUT THIRD ORDER INTERCEPT POINT =    11.0 dBm
     MINIMUM DISCERNIBLE INPUT SIGNAL   =  -139.2 dBm
     MAXIMUM SPURIOUS-FREE INPUT SIGNAL =   -55.1 dBm
     SPURIOUS-FREE DYNAMIC RANGE        =    84.1 dB
```

fig. 8. IMD analysis of a GaAs FET preamplifier.

near cutoff tends to limit its dynamic range, such that the "optimum" bias point from a noise figure standpoint often coincides with the "worst" bias point as far as dynamic range and actual system sensitivity are concerned. Remember, although we talk about desiring high "signal to noise ratio," what we really need for maximum sensitivity is a signal level that is high relative to the sum of noise and distortion. If we can considerably reduce IMD interference by giving up some slight amount of noise performance, the overall system sensitivity has to improve!

Joe Reisert, W1JR — probably the most prominent UHF DXer of our time — has long advocated designing bias circuits for preamplifiers so that device quiescent current can be readily and remotely varied.[3] This way the user can optimize noise figure when operating conditions call for it, and readily improve dynamic range, at a sacrifice in noise performance, should interference conditions dictate. Since all RF design is a series of compromises, Joe's approach seems to offer the best of all possible worlds.

There has long been controversy in Amateur circles over the relative merits of bipolar junction transistors and MOS field effect devices as VHF preamplifiers. Bipolar advocates boast the excellent low-noise performance of these devices, while those preferring the

N6TX TABLE 1
RECEIVER DYNAMIC RANGE COMPARISON

	STANDARD LEVEL DBM	HIGH LEVEL DBM	BIPOLAR JUNCTION TRANS	MOS FIELD EFFECT	GaAs FIELD EFFECT
Conversion Gain (dB)	-6	-7	12	17	24
Noise Figure (dB)	7	8	2	3	1
IMD Intercept (dBm)	+4	+22	-8	-2	+11
SSB Sensitivity (dBm)	-133	-132	-138	-137	-139
Maximum Input (dBm)	-38	-25	-60	-59	-55
Dynamic Range (dB)	95	107	78	78	84

NOTE: Measurements were performed at 144 MHz, with representative devices. Results at other frequencies will vary, but comparisons will be similar.

MOS devices cite their higher gain and stable operation, which eliminates the need for neutralization. **Figures 6** and **7** seem to indicate that neither device holds a clear advantage as far as overall system performance is concerned. The two representative amplifiers I tested in preparing this manuscript exhibited identical dynamic range.

Gallium-Arsenide Field Effect Transistors, on the other hand, are the undisputed winner in all areas of VHF and UHF performance. As indicated in fig. **8**, the GaAs FET offers exceptional high gain, low noise, and wide dynamic range performance. If only they weren't so expensive!

summary

In evaluating receiver performance, it is necessary to consider dynamic range limitations, as well as noise figure, to select the combination of devices and circuits that will yield the best overall sensitivity. **Table 1** summarizes the results of testing various competing mixer and preamplifier technologies. Although the tests were performed at 2 meters, we can generalize the results to other VHF and UHF bands as well.

It appears that best receiver performance will be achieved by cascading a GaAs FET preamplifier with a high-level doubly-balanced mixer. Two-tone analysis confirms that such a combination has considerable immunity to intermodulation and cross-modulation interference, while maintaining an impressively low-noise figure.

references

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2. H.T. Friis, *"Noise Figures of Radio Receivers," Proceedings of the IRE*, July, 1944, page 419.
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ham radio

a pulsewidth noise discriminator

Impulse noise control works on time duration rather than amplitude

I think most hams would agree that the best impulse-noise squelch would be one that stopped each noise burst at its source. Unfortunately there are too many noise bursts arriving from too many directions to make such a thing possible. It's usually difficult to locate a local source; even if you do, the person responsible for the noise is often unwilling or unable to cooperate. Once launched, these disturbances seem determined to enter our receiving systems; when they do, they're repeatedly amplified, modified, and stretched as they race from antenna terminals to speaker. In short, once admitted, these unfriendly signals are actually made worse — often *much* worse — by your receiver's own circuits.

What happens at the output after a burst of noise arrives at your receiver's input is quite predictable. Some of the fast-changing wave front is absorbed by the first tuned circuit, then released in the form of ringing at this filter's natural resonant frequency. The resulting damped oscillation is then translated by a local oscillator, resulting in a rapid rise at the input of the IF filter, which absorbs some of the energy and then releases it in the form of ringing at *its* resonant frequency . . . and so on.

basic noise control methods

Over the years a great many circuits have been tried in an effort to control impulse noise. Successful methods have been of two basic types: the noise blanker and the noise limiter. The well known Lamb filter (often called a hole-puncher, noise silencer, or noise blanker) takes a sample of each noise pulse from a receiver stage as near to the antenna as possible and, using fast circuits, forms a blanking pulse that momentarily blocks the receiver's IF stage just before the ringing pulse of noise energy arrives. The blanking pulse is designed to embrace the ringing time caused by the filter characteristic. Some rise and fall time is usually added so the blanking function will not itself generate audible clicks at the receiver's output. This system has been around for a long time; properly designed, it works very well. But one problem with this technique is that it must be designed to go into action only on noise pulses that are significantly larger than desired signals in order to avoid the creation of excessive distortion.

Another form of impulse noise control is called the "peak limiter" system. Again, this method is restricted to noise signals whose peak amplitudes are above that of desired signals. When a noise burst is received, the desired signal is momentarily suppressed and the interference is limited in peak amplitude. Perhaps the best features of this noise control system are its simplicity and its ability to reduce possible damage to our ears caused by otherwise nearly unlimited sharp audio sound transients.

Neither method is effective in removing noise bursts of low to moderate amplitude, or of durations of greater than a few microseconds, which usually includes those "woodpecker" style noise disturbances. The majority of disturbances fall in the latter category, with high amplitude disturbances in the minority.

pulsewidth noise discriminator

This article is about a third method that effectively handles a wide range of impulse noise amplitude levels and can be used either by itself or in conjunction with the more familiar methods described above. Furthermore it can be added at the *audio* output of any receiver. I have called this method the PND — Pulsewidth Noise Discriminator. Rather than working on peak amplitudes, this system makes use of time duration differences between the character of almost all desired signals and impulse noise. Impulse noise bursts at their origin exist for only a few nanoseconds to

By Don E. Hildreth, W6NRW, P.O. Box 60003, Sunnyvale, California 94088

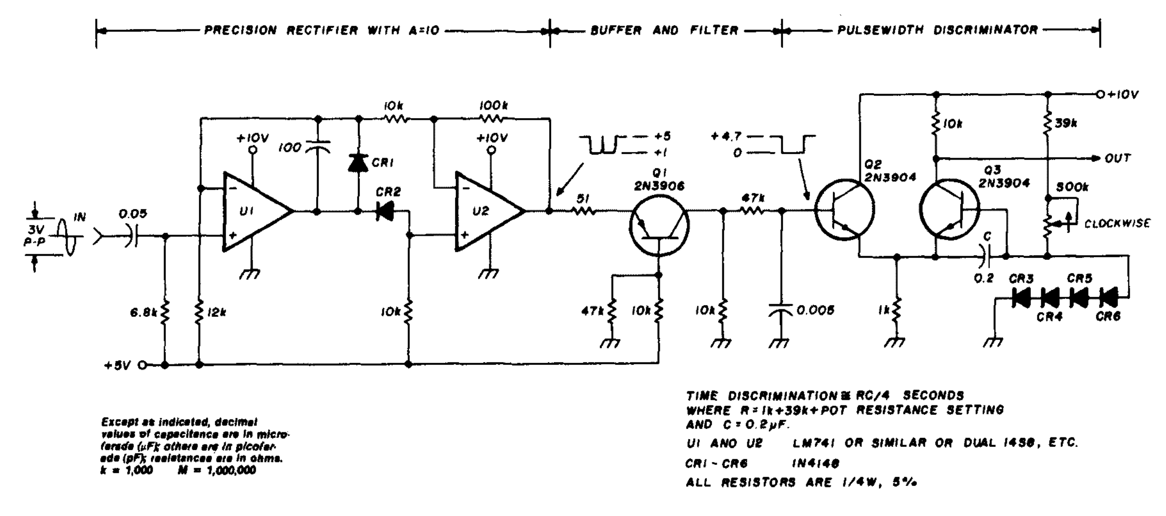


fig. 1. Basic PND system.

microseconds, but they are then transformed by receiver circuits into ringing bursts lasting from a little less than one millisecond up to as much as ten milliseconds, depending on the shape and bandwidth of the narrowest filter being used.[1] Concurrently, the vowel parts of desired SSB voice signals are fractions of a second long; even the shortest parts of CW are typically of 50 to 100 milliseconds duration for the "dit" and "space" lengths.

Using this data, a basic pulse width discriminator is designed to ignore any signal until it has existed for a selectable period of time — 10 milliseconds, for example — and will consequently block a noise pulse stretched by a typical 100 Hz bandwidth filter.

The circuit of fig. 1 is the basis of this sytem. When no signal is present, Q1 and Q2 are ON and Q3 is OFF. When a signal arrives, the precision rectifier formed by U1 and U2 develops a negative gate that turns off Q1 and Q2. However, Q3 remains off until the no-signal reverse charge of a nominal -2 volts on C, Q3's base to emitter capacitor, is bled off and reversed to the required level of approximately +0.7 volts by charging through R, which is made up of 1K + 39K + the 500K pot setting. When Q3 goes on, its collector voltage drops. It is this signal that is used to accept a desired signal or to reject those that are too short for completion of the timing cycle as determined by the effective value of R. A second and very desirable feature of the basic pulsewidth discriminator part of the circuit is that it resets very quickly, thereby avoiding the integration of noise pulses provided they are not too close together (equal to or less than the selected time discrimination period). This feature makes the full-wave rectification provided by U1 and U2 plus a small amount of filtering necessary to avoid a functional dropout between the cycles of a desired

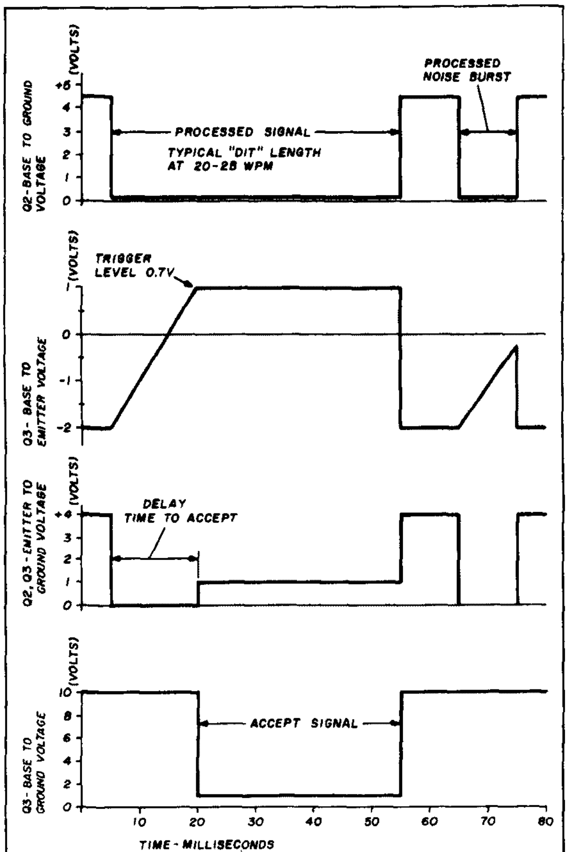


fig. 2. Idealized PND waveform.

signal. Using the specified circuit values, a minimum of 1 volt RMS is required for normal operation.

Fig. 2 shows what happens as a function of time. Since the pulsewidth discriminator circuit would ig-

nore a ringing or sinusoidal signal of any period, the incoming signal is transformed into a negative gate. An ideal gate pulse is shown for clarity. This use of an oscilloscope is convenient at the Q2 and Q3 emitter junction, to observe and measure the time discrimination performance.

a PND application

Fig. 3 shows an application in which an audio filter feeds the pulse noise discriminator and the output of the discriminator is used to key an audio white-noise generator for listening to CW. (Of course you can have it key a tone oscillator if your prefer.) The mixer control is arranged to enable listening to CW directly through the filter, or to the noise generator, or to any mixture of the two. By connecting the control in this experimental way, you can test and experience how well this idea works: simply compare a non-noise discriminated 750 Hz CW output to the processed audio-noise CW output that is driven by the PND.

In operation, the +10 volt level at Q3's collector is used to squelch noise output from the dual op-amp noise generator shown in **fig. 3** or in the SSB output of **fig. 4**. When the PND system recognizes a signal, the Q3 collector level drops, opening the transistor switch Q2 of **fig. 3** or Q1 of **fig. 4**. One undesirable feature of this noise control system relates to operating convenience: your receiver's output is normally OFF until a qualified signal appears. If you like to be aware of the noise floor, as I do, this can be a disadvantage, so I usually run the mixer control midway when looking for DX. Then, depending on conditions, I decide which way to twist this control. If I want to avoid the tinkling roar present when listening to a low-level signal through a narrow CW filter, I turn the control to admit only the keyed-noise signal with its silence between characters. But if I want to make use of the ear-brain filter capabilities (when there is more than one signal in the filter passband) then I turn the control to allow only the signal through the 750 Hz filter[2] to reach the power amplifier.

With the mixer control set toward keyed-noise operation, and with the delay control set at minimum, slowly increase the receiver gain at a no-signal spot

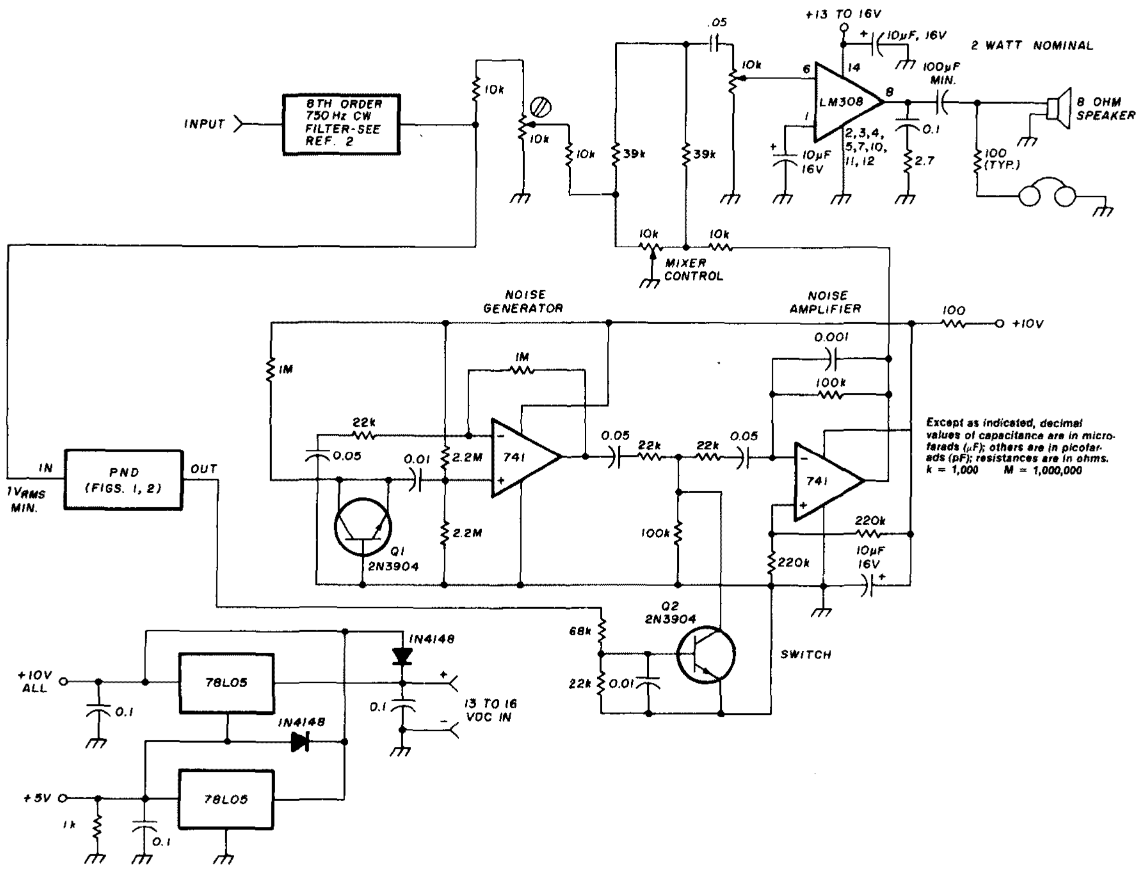
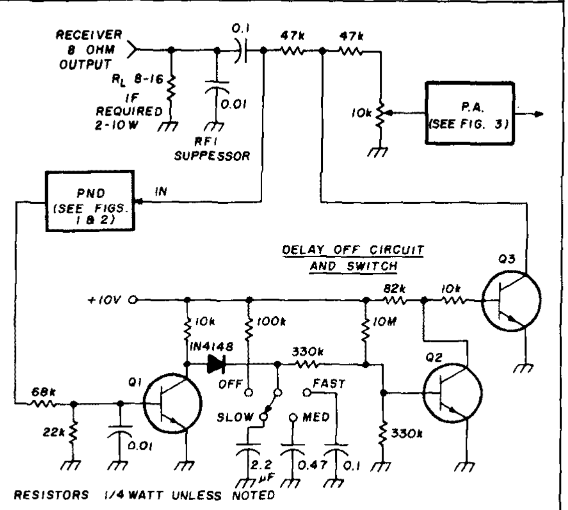


fig. 3. A PND application.



Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms.
k = 1,000   M = 1,000,000

fig. 4. Combined PND and squelch for SSB.

on the dial until the noise function starts to be heard sporadically. Increase the delay until the noise stops, then increase the delay just a bit more. This will match the delay to your filter bandwidth. If you are in a location that is too radio-quiet at the time, set the 500K pot at about mid-range for a nominal 12 millisecond delay. This will usually enable the rejection of noise pulses as they are stretched by a typical 100 Hz bandwidth filter. For best operation, the amplitude of a signal being received should be adjusted by your receiver's output level control to just a little above that required for reliable keying of the noise generator. Once this is done, the control at the output of the filter is set so that either the perceived level of keyed noise or the tone level are about the same when the mixer control is in any position. Adjust the volume control for the overall listening level desired. Once all of these settings are made, you will usually work with only your receiver's output control and the mixer control.

single sideband

Although the electronic switch as driven by the PND system can follow code quite well, its use on SSB would result in choppy voice reception. To use PND on sideband it is best to insert a delay circuit between the PND circuitry and the electronic switch to keep the controlled audio stage ON for a short time after the PND has shut off. This technique was used with my laryngeal squelch[3] and that part of the circuit is included in **fig. 4**. If you use the voice-filter system,[2] you can either use PND to open both channels, or you can feed a sample of the vowel filter to PND and control the consonant filter with the delay/switch combination. If you use the vowel filter, the delay setting is about the same as that used for CW. However, if the normal 3 kHz voice bandwidth is used the delay may usually be reduced to its minimum. The PND system, in its basic form shown here, is not as effective for SSB as it is when used to key the noise generator for CW, for although noise is rejected between a voice signal's ON times, it can appear in addition to the desired signal during the ON periods. Circuit development to improve on this problem is being studied, but requirements are much more complex.

general considerations

Although PND can handle most impulse noise problems unaided by the more prevalent noise silencers, it is still best to have a Lamb-type noise blanker in addition. Since the basic noise blanker blocks out noise bursts early in the receiver, it reduces the probability that strong pulses will drive one or several amplifier stages into heavy saturation, which can block a receiver for periods much longer than the offending noise-pulse length, and this is something that PND cannot help. Also, the front-end blanker system can suppress auto ignition pulses produced by an engine at moderately high RPMs, while PND is limited to rejecting auto ignition at idling RPMs when a 100 Hz bandwidth is used and at higher RPMs only when the bandwidth is increased. At the same time, PND can wipe out those woodpecker noise sources while their ON periods are too long for the Lamb blanker to handle. PND can be most effective when it is used in conjunction with either a blanker or a limiter, but use with a blanker is my first choice.

Because we are accustomed to using noise floor as a guide — and with PND you lose this reference — it's easy to run up your receiver gain much too high. Unfortunately, when the gain is too high, the noise-floor itself, even without an antenna, creates what amounts to a constant ringing level as it is stuffed through filters. Moreover, since the noise-floor in linear receivers is not limited, amplitude variations on this ringing can make a weird form of noise-floor-generated CW by the PND. When you hear this, just back off on the gain control a little to achieve silence.

references

1. Don E. Hildreth, W6NRW, "Graphic Filter Design," *ham radio*, April, 1984, page 38.
2. Don E. Hildreth, W6NRW, "Communications Audio Processor for Reception," *ham radio*, January, 1980, page 71.
3. Don E. Hildreth, W6NRW, "Smart Squelch," *ham radio*, June, 1983, page 37.

ham radio

A printed-circuit board (**fig. 3**) is available for $9.95 postpaid; the PC board *plus parts* (no controls) is available for $32.95 postpaid. Contact the author, Don Hildreth, W6NRW, P.O. Box 60003, Sunnyvale, California 94088.

IMD and intercept points
of cascaded stages

Use this program to determine performance parameters

Intercept point is a useful concept in predicting the spurious intermodulation products generated by components, systems, or subsystems. Only the second-order and third-order products are significant. The intercept point is the power level at which the spurious response equals the fundamental response. The value may be referred to the input or output. Usually the output is referenced for amplifiers and mixers, and the input for receivers. The intercept points for the second-order and third-order products may be the same or different, depending on the circuit of the device. Typically the responses are plotted using log-log scales with the values in dBm as shown in **figs. 1** and **2.**

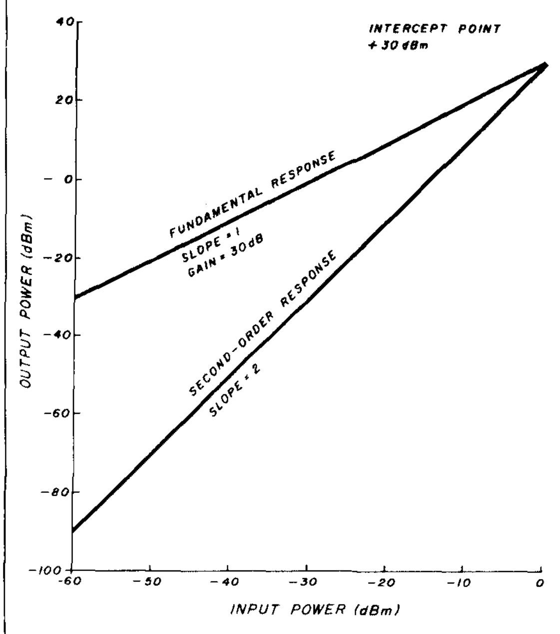


fig. 1. Typical second-order response.

Assuming two signals at frequencies f_A and f_B, and $f_A > f_B$, the second-order products are: $f_A + f_B$, $f_A - f_B$, $2f_A$, and $2f_B$. The second harmonics are not strictly intermodulation products, but may be predicted in the same manner except that their amplitudes are 6 dB less than the sum and difference products. If the two fundamental frequencies are almost equal, the $f_A - f_B$ term is near zero frequency and the remaining product is at about twice the fundamental frequencies. Half-octave filters can be used to attenuate the second-order products. Refer to **fig. 3** for the worst case with f_A and f_B at the band edges of a half-octave filter.

Third-order intermodulation products present the most serious problem for devices having bandwidths less than one-half octave. For two signals at frequencies f_A and f_B, the third-order products are: $2f_A + f_B$, $2f_A - f_B$, $2f_B + f_A$, and $2f_B - f_A$. For a narrow-band device centered at 20 MHz, two signals, $f_A = 20.50$ MHz and $f_B = 20.25$ MHz, will generate the third-order product $2f_B - f_A$ at exactly 20 MHz. For three signals at frequencies f_A, f_B, and f_C, the third-order products are: $\pm f_A \pm f_B \pm f_C$. Third-order products of three signals are seldom considered except for multi-frequency systems such as cable TV.

measurement techniques

For single or cascaded components, intercept point is measured by driving the device with two equal amplitude signals and measuring the fundamental outputs and intermodulation products on a spectrum analyzer.

The concept of intercept point for a receiver is usually limited to the RF front end. It is meaningless for the IF passband because of the nonlinearities of detection

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and gain control, as well as the high overall gain. Two equal level signals, outside the IF passband but within or as close as possible to the RF passband, are selected so that an intermodulation product is at the receiver center frequency. Their levels are simultaneously increased until an output (intermod) signal of about 10 dB signal-to-noise is observed. Record the level of the signals. Then the two signals are removed, and a single signal at the receiver center frequency is adjusted in level to produce the same output. Its level is also recorded. For third-order products the two signals are usually placed within the RF passband. However, for second-order products the signals fall out of the RF passband if the RF bandwidth is less than an octave. Refer to **fig. 3**. The most important second-order product in a receiver comes from a signal at one-half center frequency that doubles into the center frequency. This latter measurement is made by increasing the amplitude of a signal at one-half the receiver center frequency until an output signal of about 10 dB signal-to-noise is observed. Record the input level. This signal source must be well filtered so that its second harmonic is well below the second-order response. Next the input signal is tuned to the receiver center frequency and its level is adjusted to produce the same output, and this level is recorded.

When connecting two signals to the input, the insertion loss of the combiner must be subtracted from each generator output. A second precaution is to make all measurements at least 10 dB below the 1 dB compression point. Otherwise the device will be operating in its large signal area.

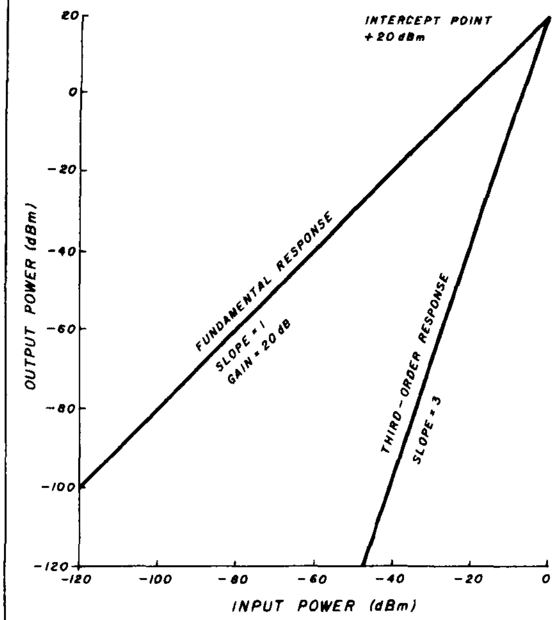


fig. 2. Typical third-order response.

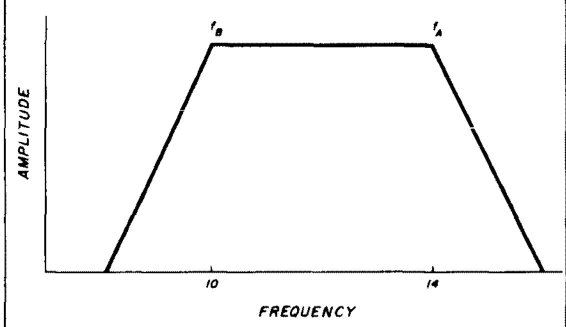


fig. 3. Half-octave filter second-order products.

second-order products

Refer to **fig. 1**. The slope of the second-order response is 2. As the fundamental output decreases by 1 dB, the second-order intermodulation products decrease by 2 dB. For two equal signals, the function may be expressed as:

$$IP = P + IMR \qquad (1)$$

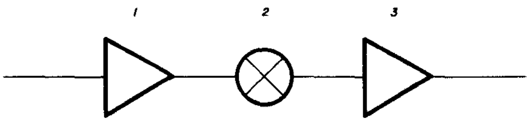
IP is the second-order intercept point in dBm, *P* is the fundamental response in dBm, and *IMR* is the ratio between the fundamental and second-order responses in dB. In the case of a receiver, *P* is the level of the two signals or half-frequency input, and *IMR* is the ratio of *P* to the level of the signal at center frequency.

For example, if the fundamental outputs of an amplifier are −10 dBm and the second-order intermodulation products are −45 dBm, the second-order intercept point is:

$$IP = -10 + 35 = +25 \ dBm$$

Knowing the intercept point, this equation will predict the second-order intermodulation products for known signal levels. For a single input signal, the same methods can be used to calculate the intercept point if the second harmonic is measured, or to predict the second harmonic if the intercept point is known. However, the second-harmonic response is 6 dB less than a second-order intermodulation product.

The levels of the second-order intermodulation products are proportional to the product of the levels of the fundamental signals. For signals of unequal levels, the level of equal signals that produce the same intermodulation products can be calculated. If the levels are in dBm, add the two and divide by one-half. For two signals, one at −20 dBm and the other at −26 dBm, the equivalent equal amplitude signals are



stage	IP dB	IP ratio	gain dB	gain ratio
1	+20	100	10	10
2	+13	20	−6	0.25
3	+30	1000	20	100

using cascaded stage equation

$$\frac{1}{\sqrt{IP}} = \frac{1}{\sqrt{IP_1 \cdot G_2 \cdot G_3}} + \frac{1}{\sqrt{IP_2 \cdot G_3}} + \frac{1}{\sqrt{IP_3}}$$

$$= \frac{1}{\sqrt{100 \cdot (0.25) \cdot 100}} + \frac{1}{\sqrt{20 \cdot 100}} + \frac{1}{\sqrt{1000}}$$

$$= 0.0200 + 0.0224 + 0.0316 = 0.074$$

$$IP = 182 = 22.6 \text{ dBm}$$

Alternate method

$$\Delta Ip = 20 \log \left[1 + \sqrt{\frac{IP_2}{IP_1 \cdot G_2}}\right]$$

For stages 1 and 2:

$$\Delta IP = 20 \log \left[1 + \sqrt{\frac{20}{100 \cdot 0.25}}\right] = 20 \log 1.89 = 5.5 \text{ dB}$$

Combining stages 1 and 2

$$IP = IP_2 - \Delta IP = 13 - 5.5 = 7.5 \text{ dBm} = 5.6 \text{ milliwatts}$$

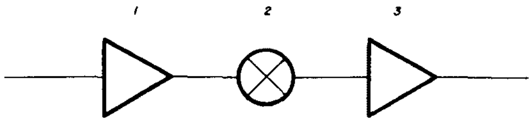
Now cascade this with stage 3

$$\Delta IP = 20 \log \left[1 + \sqrt{\frac{1000}{5.6 \cdot 100}}\right] = 20 \log 2.34 = = 7.4 \text{ dB}$$

$$IP = IP_3 - \Delta IP = 30 - 7.4 = 22.6 \text{ dBm}$$

Input intercept point = 22.6 − 34.0 = −11.4 dBm

fig. 4. Example shows two methods of calculating cascaded stages second-order intercept point.



stage	IP dB	IP ratio	gain dB	gain ratio
1	+20	100	10	10
2	+13	20	−6	0.25
3	+30	1000	20	100

using cascaded stage equation:

$$\frac{1}{IP} = \frac{1}{IP_1 \cdot G_2 \cdot G_3} + \frac{1}{IP_2 \cdot G_3} + \frac{1}{IP_3}$$

$$\frac{1}{IP} = \frac{1}{100 \cdot (0.25) \cdot 100} + \frac{1}{20 \cdot 100} + \frac{1}{1000}$$

$$= 0.00040 + 0.00050 + 0.0010 = 0.0019$$

$$IP = 526 = 27.2 \text{ dBm}$$

Alternate method

$$\Delta IP = 10 \log \left[1 + \frac{IP_2}{IP_1 \cdot G_2}\right]$$

For stages 1 and 2

$$\Delta IP = 10 \log \left[1 + \frac{20}{100 \cdot 0.25}\right] = 10 \log 1.80 = 2.6 \text{ dB}$$

Combining stages 1 and 2

$$IP = IP_2 - \Delta IP = 13 - 2.6 = 10.4 \text{ dBm} = 11.0 \text{ milliwatts}$$

Now cascade this with stage 3

$$\Delta IP = 10 \log \left[1 + \frac{1000}{11 \cdot 100}\right] = 10 \log 1.91 = 2.8 \text{ dB}$$

$$IP = IP_3 - \Delta IP = 30 - 2.8 = 27.2 \text{ dBm}$$

Input intercept point = 27.2 − 24 = 3.2 dBm

fig. 5. Example shows two methods of calculating cascaded stages third-order intercept point.

each at −23 dBm. If the levels are in milliwatts, the equivalent equal signal levels are each:

$$\sqrt{A_A \cdot A_B}$$

where A_A and A_B are individual signal levels.

If the second-order intercept point and gain of each stage are known, the overall intercept point of the cascaded stages may be found from the following formula:

$$\frac{1}{\sqrt{IP}} = \frac{1}{\sqrt{IP_1 \cdot G_2 \cdot G_3 \cdot G_4}} + \frac{1}{\sqrt{IP_2 \cdot G_3 \cdot G_4}} + \frac{1}{\sqrt{IP_3 \cdot G_4}} + \frac{1}{\sqrt{IP_4}} \qquad (2)$$

Each term of the formula has the intercept point of the stage multiplied by the gain of all of the following stages. The terms are numerical ratios, not dB or dBm. A look at each term will indicate the contribution of each stage to the overall system intercept point.

Another method of calculating the overall intercept point of cascaded stages is to use the formula:

$$\Delta IP = 20 \log \left[1 + \sqrt{\frac{IP_2}{IP_1 \cdot G_2}}\right] \qquad (3)$$

The ΔIP is in dB and is subtracted from the second stage intercept point to give the overall intercept point of the two cascaded stages. For more than two stages, this formula is used for the first two stages, and that result is then used with the third stage and so forth.

Both formulas assume the worst case in which the intermodulation products within each stage *add in phase*. If a linear stage is part of the system, it must be included with its actual gain (or loss) and an infinite intercept point.

Refer to fig. 4 for sample calculations of the intercept point of three cascaded stages. The output intercept point is calculated. The input intercept point

equals the output intercept point reduced by the total gain.

third-order products

Refer to **fig. 2**. The slope of the third-order response is 3. As the fundamental output decreases by 1 dB, the third-order intermodulation products decrease by

```
10 PRINT"INTERCEPT POINT OF CASCADED STAGES"
20 PRINT:PRINT
30 PRINT"2  SECOND-ORDER INTERMODULATION"
40 PRINT"3  THIRD-ORDER INTERMODULATION"
50 INPUT"SELECT 2 OR 3";A$
60 INPUT"ENTER NUMBER OF STAGES";C
70 CLS
80 FOR N=1 TO C
90 PRINT"ENTER INTERCEPT POINT FOR STAGE";N;"IN DBM"
100 INPUT I(N)
110 PRINT"ENTER GAIN OF STAGE";N;"IN DB"
120 INPUT G(N)
130 NEXT N
140 CLS
150 PRINT"IP(DBM)","GAIN(DB)
160 PRINT
170 FOR N=1 TO C
180 PRINT I(N),G(N)
190 NEXT N
200 PRINT:INPUT"IS DATA OK Y/N";B$
210 IF B$="N" THEN CLS
220 IF B$="N" THEN GOTO 80
230 FOR N=1 TO C
240 I(N)=10[(I(N)/10)
250 G(N)=10[(G(N)/10)
260 NEXT N
270 E(C)=1:D(C)=1/I(C):IF A$="2" THEN D(C)=SQR(D(C))
280 FOR N=(C-1) TO 1 STEP -1
290 E(N)=G(N+1)*E(N+1)
300 D(N)=I(N)*E(N)
310 D(N)=1/D(N)
320 IF A$="2" THEN D(N)=SQR(D(N))
330 NEXT N
340 FOR N=1 TO C
350 IP(N)=D(N)+IP(N-1)
360 NEXT N
370 IP(C)=1/IP(C)
380 IP(C)=10*LOG(IP(C))/LOG(10):PRINT
390 IF A$="2" THEN IP(C)=2*IP(C)
400 IF A$="2" THEN PRINT "SECOND-ORDER INTERCEPT POINT IS ";IP(C);" DBM"
410 IF A$="3" THEN PRINT "THIRD-ORDER INTERCEPT POINT IS ";IP(C);" DBM"
```

fig. 6. TRS-80 Model III program listing determines the intercept point of cascaded stages.

```
INTERCEPT POINT OF CASCADED STAGES


2  SECOND-ORDER INTERMODULATION
3  THIRD-ORDER INTERMODULATION
SELECT 2 OR 3? 3
ENTER NUMBER OF STAGES? 3
ENTER INTERCEPT POINT FOR STAGE 1 IN DBM
? 20
ENTER GAIN OF STAGE 1 IN DB
? 10
ENTER INTERCEPT POINT FOR STAGE 2 IN DBM
? 13
ENTER GAIN OF STAGE 2 IN DB
? -6
ENTER INTERCEPT POINT FOR STAGE 3 IN DBM
? 30
ENTER GAIN OF STAGE 3 IN DB
? 20
IP(DBM)              GAIN(DB)

 20                   10
 13                  -6
 30                   20

IS DATA OK Y/N? Y

THIRD-ORDER INTERCEPT POINT IS  27.2141   DBM
```

fig. 7. Three-stage device IMD intercept point calculation is simple with user-friendly program.

3 dB. For equal signals, the curve may be expressed as:

$$IP = P + 1/2(IMR) \qquad \textbf{(4)}$$

IP is the third-order intercept point in dBm, and IMR is the ratio between the fundamental and third-order responses in dB. For the case of a receiver, *P* is the level of the two input signals and *IMR* is the ratio of *P* to the level of the signal at center frequency.

For example, if the fundamental outputs of an amplifier are −10 dBm and the third-order intermodulation products are −50 dBm, the third order intercept point is:

$$IP = -10 + 1/2(40) = +10\ dBm$$

Knowing the intercept point, the equation will predict the third-order intermodulation products for known signal levels.

The levels of the third-order intermodulation products are proportional to (1) the cube root of the product of three signals or (2) in the case of two signals, the cube root of the square of the higher level signal times the other. For signals of unequal levels, the equivalent equal level signals that produce the same intermodulation products can be calculated. If the levels are in dBm, for the two signals add 2/3 of the larger to 1/3 the smaller. If one signal is at −20 dBm and the other at −32 dBm, the equivalent equal level signals are at −24 dBm. For three signals, add 1/3 of each level in dBm. If the levels are in milliwatts, the equivalent levels are

$$\sqrt[3]{A_A \cdot A_B \cdot A_C} \text{ or } \sqrt[3]{A^2{}_A \cdot A_B}$$

where A_A is the highest level.

If the third-order intercept point and gain of each stage are known, the overall intercept point of cascaded stages may be found from the following formula:

$$\frac{1}{IP} = \frac{1}{IP_1 \cdot G_2 \cdot G_3 \cdot G_4} + \frac{1}{IP_2 \cdot G_3 \cdot G_4} + \frac{1}{IP_3 \cdot G_4} + \frac{1}{IP_4} \qquad \textbf{(4)}$$

Each term of the formula has the intercept point of the stage multiplied by the gain of all the following stages. The terms are numerical ratios, not dB or dBm. A look at each term indicates the contribution of each stage to the overall system intercept point.

Another method of calculating the overall intercept point of cascaded stages is to use the formula:*

$$\Delta IP = 10 \log \left[1 + \frac{IP_2}{IP_1 \cdot G_2}\right] \qquad \textbf{(5)}$$

The ΔIP is in dB and is subtracted from the second-

*Note differences between **eqs. 2** and **4** and **3** and **5**. — **Editor.**

stage intercept point to give the overall intercept point of the two cascaded stages. For more than two stages, this formula is used for the first two stages, and the result is then used with the third stage and so forth.

Both formulas assume the worst case in which the intermodulation products within each stage add in phase. If a linear stage is part of the system, it must be included with its actual gain (or loss) and an infinite intercept point.

Refer to **fig. 5** for a sample calculation of third-order intercept point for three cascaded stages. The output intercept point is calculated. The input intercept point equals the output intercept point reduced by the total gain.

computer program aids calculation

Figure 6 lists the steps of a typical BASIC language computer program for calculating the second-order and third-order intercept points of cascaded stages if the values for the individual stages are known.

Figure 7 is a TRS-80 Model III™ printout showing a typical calculation of third-order IMD intercept point of a three-stage device.

ham radio

the Russian Woodpecker: a continuing nuisance

What it is and what can be done

It never fails . . . you're working some choice DX on 20 meters for your 300th country or you're in one of the Area traffic nets, trying to pass a message to another ham a thousand miles away. Suddenly, without warning, the band is shattered by something that sounds like a cross between a machine gun and a jackhammer. No, it's not the neighbor's power saw or the family microwave oven . . . it's the Russian Woodpecker in full operation. With the interference level running 10 to 20 dB over S9, your much-needed contact is buried under this avalanche of QRM and heard no more. The only solution is to turn off the rig and cool down with a tall, cold 807.

What is the Woodpecker? Why is it in operation? And why does the Soviet Union persist in creating this level of interference worldwide? What can we do about it? And what have other radio services and users of the spectrum done? This article will explore the aspects of this problem and suggest some possible solutions.

Basically, the Russian Woodpecker is an extremely powerful over-the-horizon radar system. It operates over most of the HF band, with effective radiated power levels of some 10 to 50 megawatts. To understand the nature of this problem, we need to have a basic understanding of how OTHB (Over-The-Horizon-Backscatter) radar operates, some sense of the history of experimentation and operation in this field, and an educated awareness of Soviet diplomatic response to complaints about the interference their system is generating.

basic radar operation

It has long been recognized that radar can be operated on any frequency. The earliest radar systems — built by the British and responsible for much of that nation's success during the Battle of Britain — were operated at a frequency around 30 MHz. This was due, in part, to 30 MHz being the highest frequency at which significant levels of power could be generated. Later radar systems were operated at much higher frequencies as technology developed tubes capable of generating multi-kilowatt levels of power at shorter and shorter wavelengths. Moving through the spectrum from VHF to UHF and finally into the microwave regions, radar engineers have traditionally sought the highest possible frequency of operation for several reasons. Shorter wavelengths bring increased target resolution and give the system, as a whole, increased immunity to interference, either natural or man-made. In addition, highly directional antennas become physically smaller, making possible the design of mobile radar units with greater target discrimination.

Unfortunately, all these radar systems suffer from a single common defect: they can operate only on line-of-sight. This means that at greater distances, the target must be at higher altitudes in order to be detected. Aircraft or cruise missiles flying at very low altitudes can escape radar detection

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until they are almost on top of the radar transmitter. Thus, several aircraft flying at tree-top level could approach and neutralize radar installations undetected, leaving a blind spot through which an enemy could pour aircraft or missiles. This scenario, dealing with the problem of low level detection, has left many a defense planner, both American and Soviet, in a cold sweat.

lower frequencies provided new opportunities

It had been recognized by many that some form of high-frequency radar, utilizing backscatter techniques, could detect these low-level targets. Since the radar signal would be reflected off the ionosphere and illuminate the target from above, there would be no escape from this type of detection system. It was also recognized that there were several inherent problems in this approach. First, the ionosphere was thought to be in a state of continuous flux, unable to provide stable refraction characteristics for any length of time; second, there would be continuous interference both to and from other users of the HF spectrum; and, finally, the reception of backscattered signals would require extremely complex detection systems.

By the early 1970's, scientific inquiry and experiments brought new light to this gloomy picture. The widespread use of ionospheric sounders, both ground-based and satellite, had shown the ionosphere to be more stable than previously thought. It was discovered that the refractive characteristics of the ionosphere changed very little in the short term — that is, for periods of approximately 30 minutes, the ionosphere is remarkably homogeneous. During the course of a day, these characteristics change in response to shifts of solar flux and geomagnetic activity. This meant a radar system would have to be capable of operating over much of the HF band to provide coverage of selected areas. Simply put, the radar would have to be frequency-agile to follow these changes in the Maximum Usable Frequency (MUF).

The explosion of computer technology made possible the correlation and analysis of weak backscattered signals on a real-time basis. Using cross-correlation reception techniques coupled with the development of magnetic drums for data storage, high-speed computers were used to sort out interference in the system. These computers could not only discern a weak target signal from ground clutter but also selectively filter out other users of the HF spectrum.

early OTHBs

In the 1950's, the United States Naval Research Laboratory and other groups began small-scale experiments with OTHB radar. These early experiments led to the solution of some of the major problems in designing a functional HF radar. Among these problems were the following:

- The return from prospective targets would be 40 to 80 dB weaker than the ground return (i.e., ground clutter).
- It was not known whether sufficient angular resolution could be developed at HF wavelengths to permit accurate target identification.
- Extremely precise doppler techniques would have to be used to permit identification.

The magnitude of this doppler problem may be seen in the following equation:

$$fd = \frac{2\, V_r f_o}{C}$$

fd represents the doppler shift, f_o represents the radar carrier frequency, V_r is the target relative velocity, and C is the speed of light. For aircraft type targets, the doppler shift varies from tenths of a Hertz upward to 50 Hz. This is dependent upon the operating frequency. The development of technologies to deal with these and other problems have resulted in the operation of both American and Soviet OTHB radar systems.

Both the American CONUS OTH-B (Continental United States Over-The-Horizon Backscatter — see sidebar, page 43) and the Soviet Woodpecker share characteristics common to all HF radars. The interaction of these characteristics may be seen from an examination of the radar range equation:

$$R_{MAX} = \frac{P_{AV}\, G_T\, G_R\, \lambda^2\, \sigma\, F_P\, T_C}{(4\pi)^3\, N_o\, (S/N)\, L_S}$$

where R_{MAX} = maximum range
P_{AV} = average power
G_T = gain of transmitting antenna
G_R = gain of receiving antenna
λ = wavelength
σ = target cross section
F_P = propagation effects factor
T_C = coherent processing time
N_o = noise power/unit bandwidth
S/N = signal to noise ratio required for detection
L_S = system losses

The major differences between HF and microwave radar systems are related to the following:

- Propagation effects — energy loss over ionospheric paths, polarization mismatch between transmitted and received signals, and gains or losses related to the dynamic nature of the transmission path.
- The amount of noise injected into the system by natural sources (i.e., distant thunderstorms) and, more importantly, by other users of the HF spectrum

(e.g., international broadcasting, Amateurs, Maritime mobile, etc.).

• Processing time (the number of hits integrated divided by the pulse-repetition frequency) — important since doppler radar requires a dwell time of T_C seconds to realize a frequency resolution of $1/T_C$ Hertz.

The transmitted waveform for HF radar systems is similar to that of microwave systems. It can be CW, pulse, FM-CW, or some other coded mode of transmission. OTHB radar have different problems with detection at minimum ranges than do microwave radars. This is because of the existence of a skip zone — that region, familiar to all hams, from which no signal is received. This skip zone accounts for HF radar's inability to detect targets closer than 1000 km to the transmitter.

A long pulse is used in HF radar to increase the sensitivity of the system and may reduce to interference levels associated with pulse modulation. In addition, the pulse repetition frequency is normally low to avoid range ambiguities. A PRF of 50 Hz will yield an unambigious range of some 3000 km. Individual pulse widths may range from tens of microseconds to several milliseconds depending upon the sensitivity desired and the desire to reduce interference to other services.

antenna requirements are severe

OTHB radar places more demands upon the antenna system than do other types of radar. The antenna must be physically large because of the low frequencies involved, be capable of handling very large amounts of power, exhibit gain and directivity over a wide range of frequencies, and be steerable in both elevation and azimuth. Typically the antenna consists of phased arrays of vertical bowtie driven elements in front of screen reflectors. The antenna lobes are steered in azimuth and elevation by shifting the phase relationships between individual active elements. Normally, separate antennas are used for transmitting and receiving. While this increases the problems of synchronizing the transmitter with the receiver, it is more than offset by the simplification of antenna construction. To place the first lobe as near horiziontal as possible, an extended ground screen is placed in front of the array. This ground screen may extend up to 3000 meters in front of and be as wide as the antenna array.

Changes in the ionosphere bring about changes in the MUF. HF radar adapts to these changing conditions by shifting its operating frequency. The ionosphere is probed with a sounder and the profiles are updated constantly. This gives real-time information as to what the best operating frequency for coverage of a particular area of interest will be. The relationship between vertical profiles and transmission paths can be seen from **figs. 1** and **2**. As the transmission frequency approaches the MUF, the paths lengthen, providing the maximum distance in a single hop transmission. Operating at or near the MUF greatly reduces path losses. Since these radar systems are not limited to a few discrete bands of frequencies, as are other services (including hams), they can follow the MUF quite closely.

The reliability of HF radar is related to antenna size, radiated power, and the range of frequencies used. These factors can overcome shortcomings in the reliability of the ionospheric paths. The ionosphere places limits on operation in both summer and winter, but for different reasons. In summer, ionization extends well into the lower regions, which normally contain neutral particles. Thus strengthened, the D-layer causes increased path loss. During the winter, decreased solar radiation creates lower electron densities in the F-layer and results in lower frequencies being required for reliable transmission. Several other problems exist because of changes within the ionosphere. These problems include the following:

• Propagation velocity is frequency dependent which places lower limits on pulse length and range resolution.

• The refractive characteristics of the ionosphere allow specific areas to be covered only by a narrow range of frequencies at any given moment.

the solar jammer

At frequencies in the high HF and low VHF range, natural extra-terrestrial sources of interference can play havoc with radar systems. During the height of the Battle of Britain, for example, British radars operating around 30 MHz were suddenly jammed by a strange, unknown signal. The interference became so severe that the British High Command felt sure it was some new and very effective form of German jamming. A group of engineers and astronomers, led by Stanley Hey, was detailed to locate the source and develop countermeasures. Together they determined that the interference appeared to originate in the area of the Sun. After photographs revealed a large sunspot group on the surface of the Sun, Hey concluded that the intensity of interference was related to the size and position of the sunspots on the solar surface. This discovery, confirmed by other investigators, led to the post-war development of solar radio astronomy.

table 1. **Current capabilities of United States and USSR OTHB radars.**

range coverage:	1000-4000 km, with longer ranges possible through multiple-hop transmissions.
angle coverage:	360 degrees in azimuth possible, but less than 120 degrees typical.
range resolution:	As low as 2 km, with 20-40 km typical.
absolute range accuracy:	10-20 km, assuming accurate and timely assessments of the ionosphere and optimum operating frequency.
angle resolution:	Determined by beamwidth. It can be as low as 1 degree, which corresponds to 50 km at a range of 3000 km.
Doppler resolution:	Generally resolution of 1/10 Hz is possible. At an operating frequency of 20 MHz it corresponds to a target velocity of just under 2 MPH.
level of interference:	Dependent upon such factors as frequency of operation, antenna design, power level, type of modulation, area of illumination, and duration of operation at any particular frequency.

• The propagation medium is filled with unwanted clutter from meteor and auroral ionization in addition to other areas of scattering that compete with target returns.

present OTHBs

The current capabilities of OTHB radars, both American and Soviet, are shown in **table 1**.

it all started with Ivan The Terrible

The initial evidence of Soviet OTHB radar capability surfaced in mid-1976. The first of these radar units, nicknamed "The Kiev Buzzsaw" or "Ivan The Terrible" was a 2-million watt transmitter operating near the city of Kiev, augmented by a smaller installation near the Black Sea town of Nikolayev. From these initial efforts, the Soviets have expanded their system into a fully functional early-warning high-frequency radar. Most of the early information concerning the Russian Woodpecker, as it is now known, came from the worldwide efforts of Amateur Radio operators. Even today, little hard information is available concerning the physical make-up of these radar installations. Western intelligence reports remain classified and, or course, the Russians appear reluctant to volunteer anything.

The Woodpecker is part of some 7000 surveillance radar systems deployed at over 1200 sites across the length and breadth of the USSR. While it was initially thought that the Woodpecker was designed for aircraft or ship detection, recent information indicates that it is actually a ballistic missile early warning system. There are currently three of these OTHB radars in operation. Two of them pointed at the United States and the other was directed at central China. These radar systems operate in conjunction with satellite detection systems to provide upwards of 30 minutes warning of an ICBM strike launched from sites within the United States or China. This HF radar launch detection system is not as accurate or reliable as a satellite system, but the two working together give 24 hour-a-day coverage of missile silos.

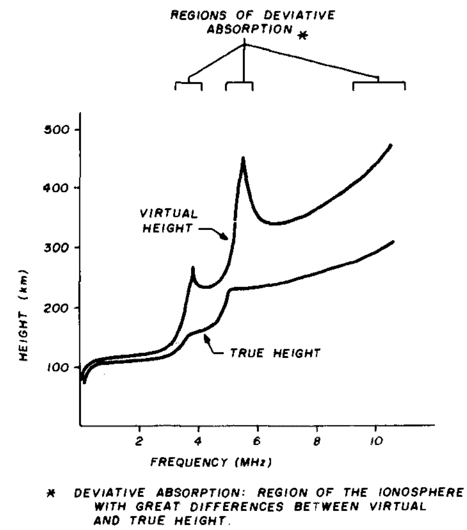

fig. 1. Daytime vertical profile of the ionosphere.*

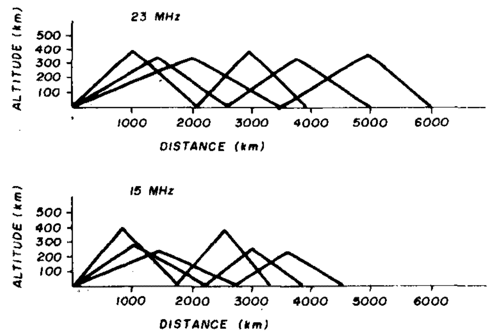

fig. 2. Virtual path traces at two frequencies based on data in *fig. 1*. Radiation angles of 4-12-20 degrees above horizontal. (Adapted from "Over-The-Horizon Radar on the HF Band," *Proceedings of the IEEE*, June, 1974.)

*Adapted, with permission, from James M. Headrick's "Over-The-Horizon Radar on the HF Band," *Proceedings of the IEEE*, Volume 62, Nov. 6, June, 1974. ©1974, IEEE.

Since its beginning in 1976, the Soviet OTHB systems have increased their power and currently operate at the 20 to 50 megawatt level. Their system utilizes pulse modulation, in contrast to the American CONUS OTHB, which transmits FM-CW. The PRF (Pulse Repetition Frequency) is normally 10 per second, although additional analysis has suggested each pulse actually consists of a pulse train of up to twenty different square wave pulses with some less than two milliseconds long, giving an effective PRF of 800 pps. The modulation scheme employed by the Russians has undergone some evolutionary changes since the inception of this radar system. Currently, the modulation, though still a pulse system, causes the radar signal to be spread in frequency. This permits frequency compression on the receiving end and results in "processing gain" for the system as a whole. In addition, this spread-spectrum technique allows the detection system to more easily discriminate against other stations on frequency. Unfortunately, these wide-band signals have further increased the interference levels to other, legitimate users of the HF band.

Currently, the radar signals no longer sit on one frequency for extended periods of time, as they once did. This is due, in part, to the protests of other users of the HF spectrum, but also to the Soviets' efforts to utilize the optimum transmission frequency. At the present time, the signals move up and down the band in 100 kHz steps at intervals of 30 seconds to 10 minutes.

why hams are most affected

It is also noteworthy that the Woodpecker chooses parts of the HF spectrum with low rates of RF occupancy. Certain portions of the band have few users per unit time and those users operate with low levels of radiated power. These areas of the spectrum are a natural for radar operation, placing less stringent requirements on the detection system. As can be seen in **fig. 3**, the Amateur bands fit nicely into this category. This helps to explain why hams have suffered the most. In addition, Amateurs tend to have limited political "pull" with their governments and, thus, are less able to bring pressure to bear to curb this interference than are other users of the spectrum. Other services, such as international broadcasting, can overcome the Woodpecker by raising their effective radiated power into the megawatt range and thereby swamping out the Russian radar.

Worldwide response to the Woodpecker arose almost immediately after its first transmission. In July, 1976, the Federal Communications Commission sent a telegram — prompted by complaints from ham operators about interference levels on the 20-meter

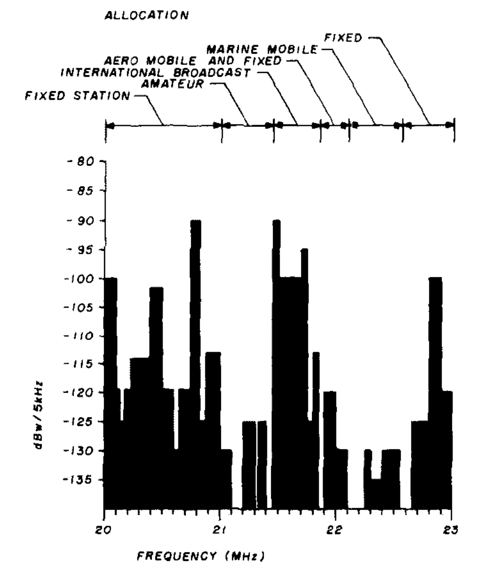

fig. 3. Typical power spectrum at HF. Note correlation with user allocation. Noise floor — 140 dBw (140 dB below 1 watt). (Adapted from "Over-The-Horizon Radar on the HF Band," *Proceedings of the IEEE*, June, 1974.)

band — to the Soviet Ministry of Post and Telecommunications. With no response from the USSR, the FCC sent three more cables. Still there was no response. In October, the FCC filed a formal complaint with the International Frequency Registration Board.

Additional complaints poured in from Amateur, Maritime, and aeronautical operators in other countries. In addition to the United States, and European nations, countries in the region of the Baltic Sea as well as Australia and New Zealand voiced strong protests. Early in 1977, the Soviet Union admitted that their experiments might cause some interference to other radio facilities for short periods of time. As worldwide pressure mounted, the USSR agreed to cut back on these radar transmissions. In reality, the Woodpecker remained on the air for the same amount of time, but its signals moved back and forth through the HF band rather than staying in one spot for extended periods of time.

In 1979, this issue surfaced, but was never pressed, at the World Administrative Radio Conference. In retrospect, this was probably for the best. This conference resulted in substantial gains for the Amateur community that might never have come about if the Conference had been disrupted over the Woodpecker issue.

Soviets ignore treaty

The USSR is signatory to international telecom-

munications treaties that spell out, in detail, the allocations for broadcasting. However, the Soviets have made full use of an escape clause included in all of these treaties. Simply put, a nation may ignore the treaty if such action is deemed to be in the best interests of its national defense. In addition, telecommunications treaties are only as good as a nation's willingness to abide by them. There is no practical way to force compliance by other countries. Most nations observe these treaties rather closely, however, realizing that compliance is in the best interest of the world community.

The current position of the United States was recently stated by Dr. William Schneider, Under Secretary of State for Security Assistance, Science and Technology. In an interview, Schneider commented, "We are making every effort to encourage the USSR to comply with their treaty obligations. In this regard, I hope we will be more effective in the future than we have been in the past."[1] In reality, this means that the Soviets will continue to use the Woodpecker as long as it suits their needs or until they develop a completely accurate and reliable satellite surveillance system for ICBM launch detection.

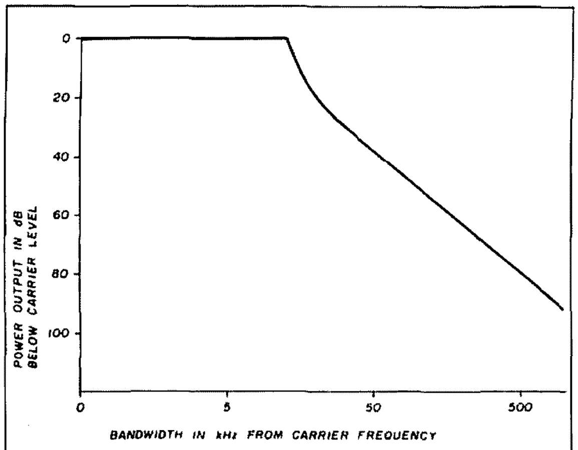

fig. 4. Spectral distribution of signal from the Russian Woodpecker (assuming 50 Mw carrier ERP, note that the Woodpecker still has 5000 watts ERP 50 kHz either side of center frequency).

what can we do?

So what can you do the next time the Woodpecker blows the 20-meter band apart? There's no point in complaining to the FCC or Department of State — they're not interested. They have literally *thousands* of complaints on file and don't need any more. They are fully aware of the problem and realize how little they can do to change it. Cranking your keyer up to 99 WPM and shooting a string of pulses in the direction of the Soviet Union is equally futile. Because the radar is designed to ignore this type of interference, all this accomplishes is additional QRM for other hams.

Perhaps the best solution to Woodpecker interference lies in the field of electronics. The technology is available to eliminate this pest at the receiver. The newer transceivers, such as the Kenwood TS-930, the ICOM IC-751, and the Yaesu FT-1, among others, have dual noise blankers, one of which is designed to eliminate long pulse noise such as that from the Woodpecker. This trend is likely to continue until most new rigs have this capability.

All this doesn't help those of us who aren't quite ready to buy a new state-of-the-art transceiver. What can we do? We have two choices. The first is to build a custom noise blanker for our existing rigs. Circuits to eliminate the Woodpecker have been published in Amateur magazines and in the ARRL's *Radio Amateur's Handbook*. The second choice is the purchase of a "Moscow Muffler," a Woodpecker noise blanker manufactured by AEA (Advanced Electronics Applications) of Lynnwood, Washington. Installed between the transceiver and antenna, this unit effectively blanks out the Woodpecker by means of an RF sensing unit that automatically takes it out of the circuit when the transmitter is keyed. The blanking width and synchronization are both adjustable. The basic sync rate may be switched between 10 and 16 Hz to allow for blanking when both OTHB radars are in operation.

It does not appear that the Woodpecker will dis-

interference not inevitable

The USA's CONUS OTH-B radar has received widespread publicity in technical, professional, and Amateur publications. At the onset of operation, the project's organizers actually *solicited* interference reports from all users of the HF spectrum. A committee was set up to handle the expected deluge of complaints; after two years of operation, only eight reports had been received. Of these, seven were disallowed because the radar had not been in operation at the time the alleged interference occurred or because the radar was operating on a frequency far removed from the one specified in the complaint. The eighth report was not a complaint at all, but rather a report from an SWL looking for confirmation. This absence of interference to other services is due to the nature of the American radar and the care exercised in the selection of clear frequencies.

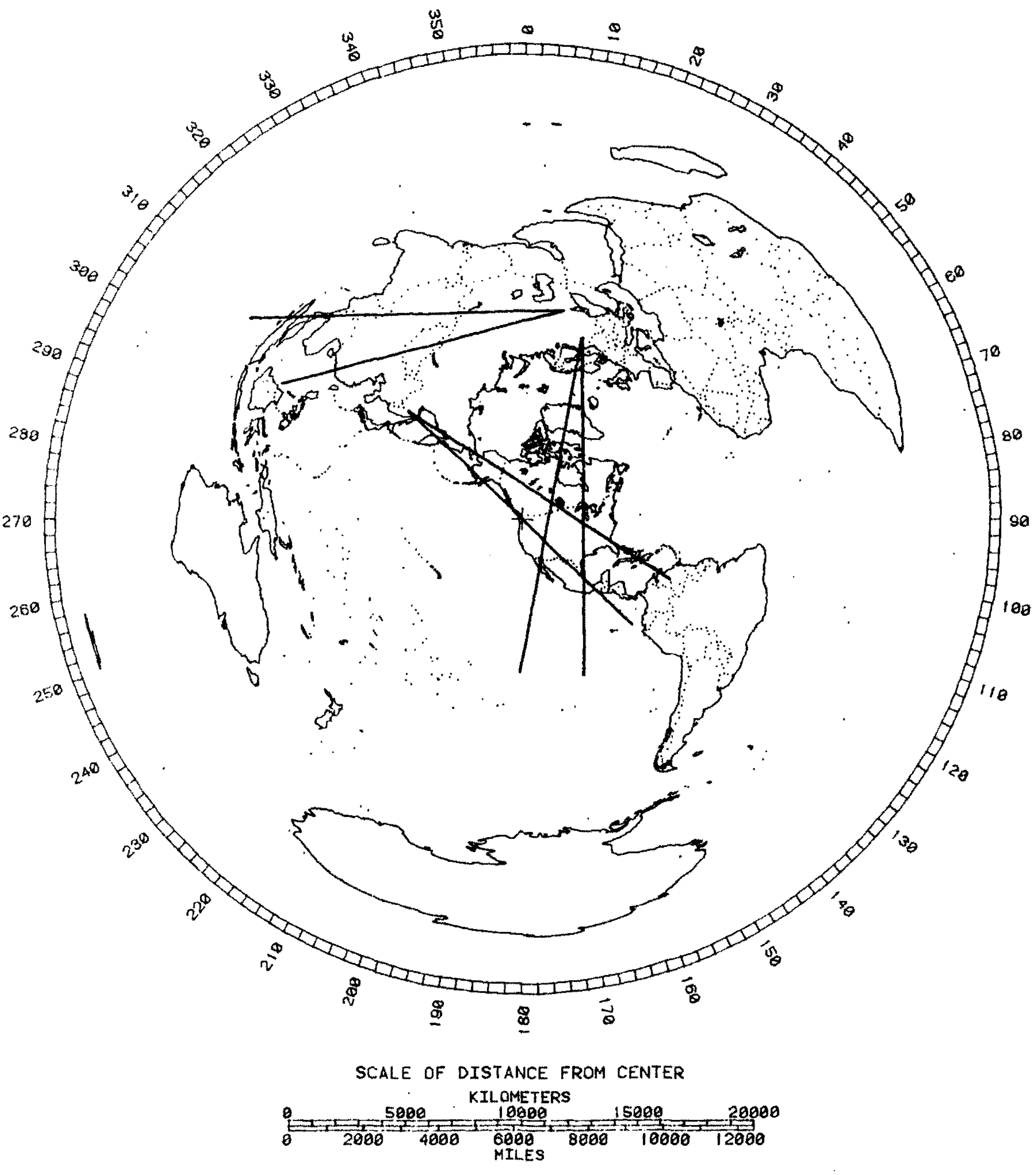

fig. 5. Coverage of Soviet "Woodpecker" ICBM launch detection systems. (Note: this base map appeared in KR7L's "Fundamentals of Grayline Propagation," *ham radio,* August, 1984, page 77. Azimuthal-equidistant map prepared by Bill Johnston, N5KR, 1808 Pomona Drive, Las Cruces, New Mexico 88001.)

appear within the near future. The Soviet Union will continue operation despite world opinion, as long as it deems the practice necessary. The ultimate practical solution will be the inclusion into Amateur equipment of noise blankers capable of removing this interference. Advancing electronic technology will provide the solution to a worldwide problem that apparently cannot be resolved by diplomatic methods.

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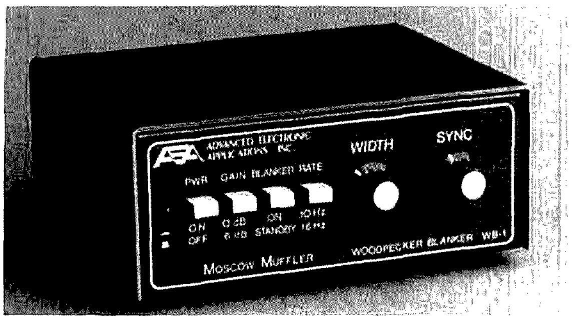

fig. 6. The MOSCOW MUFFLER® by Advanced Electronics Applications, Inc. This transceiver accessory is capable of removing interference from the Woodpecker.

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ham radio

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(See "Publisher's Log," April, 1984, page 6, for details.)

603-878-1441

a double conversion
portable SW receiver

Modify an earlier design for additional coverage, built-in frequency counter

In the April, 1984 issue of *ham radio*, Jack described the construction of a compact, portable, high performance shortwave receiver for the 46- through 100-meter bands.[1] This article provides a design for an extended coverage receiver based on that design, but offering front-end RF tuning and a built-in frequency counter and power supply. Helpful circuit hints applicable to other receiver designs are also described. The April article should be reviewed for schematics, component values and contruction details. Figures 1, 2, and 3 show several views of the new receiver from different angles, including component layout and shielding requirements — Editor.

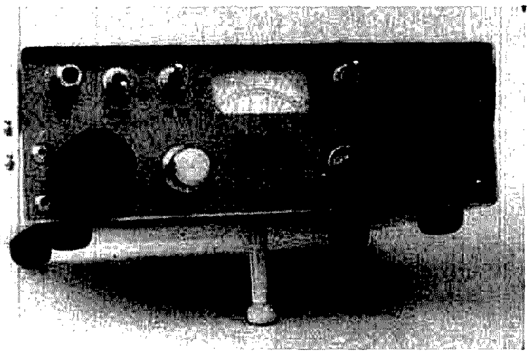

fig. 1. Top row of controls includes antenna input, on/off switch, RF bandswitching. Bottom row includes AF gain knob, RF tuning, main tuning.

Good converter design calls for an examination of all mixing by-products for each choice of local oscillator and desired input signal range and minimizes in-band spurious responses.[2] The frequency conversion scheme finally decided upon extends front-end coverage to include signals in the 9.3 to 10 MHz range. The incoming signal is downconverted to 3.3 to 4.0 MHz using a crystal oscillator and active mixer. The digital display is made to "track" by converting the "3" MHz readout to "9" MHz simply by switching the "F" LED segment, thereby eliminating the need for elaborate frequency readout conversion schemes. To accomplish this a 4-pole, 2-position C&K miniature switch performs the following functions:

- supplies +12 VDC to the converter board
- bandswitches the converter input
- bandswitches the converter output
- switches the "F" LED between "3" and "9"

The converter has been designed for a broadband response and the RF and MIX trimmers should be stagger-tuned for flatest front-end response. The converter schematics and the wiring of the CONV bandswitch are detailed in **fig. 4**.

construction details

In addition to the schematics and photos shown, the following information should be useful.

Power transformer. This should supply 14 volts at 120 mA. A 15-volt unit would probably be better to use because it would deliver (under load) a DC voltage closer to that of a car battery. Presently, power drain

By Jack Perolo, PY2PE1C, P.O. Box 2390, Sao Paolo, Brazil

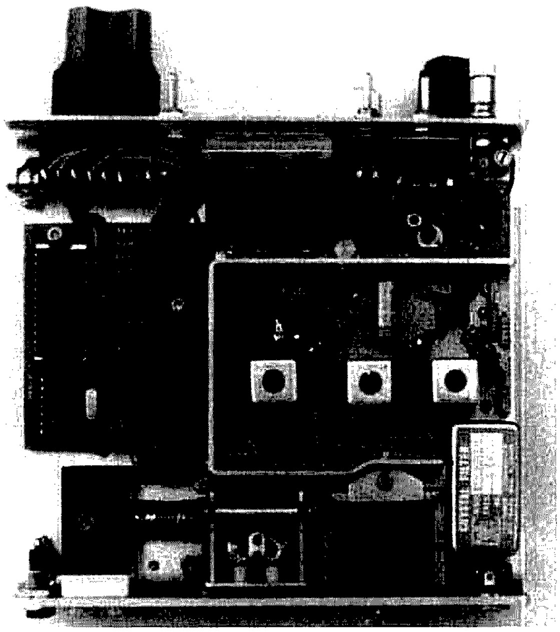

fig. 2. Top view. At top left is the 4-digit display, wired to the frequency counter below it. At top center is the S-meter, with C&K switches to its right. The input 9 MHz trap is at top right. The IF strip is below the S-meter, shielded with 1/16 inch aluminum sheet. At bottom left is the 60:1 ratio Muffett gear reducer connected with flexible coupling to the 104 pF variable capacitor. To the right of the variable capacitor is the power transformer, followed by the 9 MHz crystal filter. The back panel has provisions for two AF output jacks, DC power (12 volt) input jacks, and a 110 VAC connector (see fig. 3).

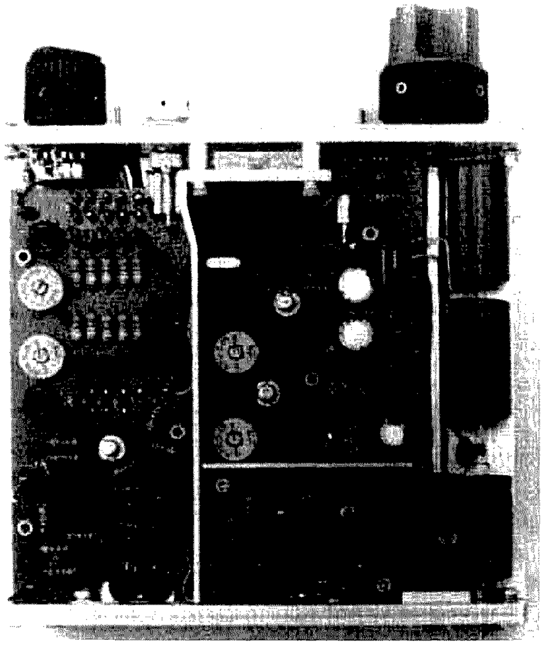

fig. 3. Bottom view. At top left are the audio frequency potentiometer and 5-24 pF Jackson RF trimmer. S-meter is at top center; below the S-meter is the PC board for power supply, AF strip and the front-end converter. The power supply electrolytic capacitors are at its right. At left is the RF/Mixer PC; notice shield between it and the front-end converter. The 110 VAC connector is at bottom. At bottom right is the VFO PC board, with electronic bandswitching circuitry, shielded from both the RF PC and front end converter PC; the VFO PC board ends near the gear reducer, seen at bottom right. All RF transistors are mounted in sockets to ease replacement in case of failure.

is 120-125 mA with the 9 MHz converter on, dropping to about 100-105 mA with the converter off. (Saw off the original brackets; use a bolted pillar and pressure holder to keep it in place.)

Space-saving techniques. In order to make room for the converter and power transformer, a new layout was developed. The audio strip PC board was redesigned and reduced in size, with all components vertically mounted. This reduced it from 1.57 x 2 inches (40 x 50 mm) to 1.18 x 1.57 inches (30 x 40 mm).

The PC board that houses the audio strip also includes the power supply, the zener diodes, and the front-end converter. The 1/16 inch thick epoxy board measures 2-3/8 x 3-1/2 inches (60 x 90 mm). Separate diodes were used in the supply to avoid confusion in case 110 VAC and external power were left on at the same time.

Gear reducer. Zero backlash, Muffett size 1 with gear ratio 60:1. Available in the U.K. for $75.00.

Cabinet. 2 x 6 x 6 inches (50 x 150 x 150 mm) HWD.

ham band options

The 80-meter band, covered in the earlier project, is included in the current version. The 40-meter band can be covered by the direct method or by the converter method. Using the direct method extends the coverage of the basic receiver to 7 MHz. (The MHz digit of the display must be read.) The converter option would employ a 4 MHz crystal; the digit problem appears to be easier to solve, but some spurious signals are likely to appear within the band. Coverage on 30 and 20 meters can be implemented by modifying the front-end conversion using a single set of coils and electronically bandswitching the parallel capacitors and the oscillator crystal. One can cover 7, 10, and 14 MHz with the same basic converter by increasing PC board size slightly.

The frequency counter can easily handle the fifth digit (tens of MHz) because the 7207 IC has provisions for it, but the power supply must be sized accordingly for the extra load. Care should also be taken in the layout and design of the front panel, which is presently

Coils needed for the converter are L14 through L16.

coil	description	frequency range (MHz)
L14,L15	RF coil 3 turns and 20 turns (No. 26 enamel wire), respectively on Amidon T-37-2 toroid	9.0-10.0
L16	RF coil 20 turns (No. 26 enamel wire) on Amidon T-37-2 toroid	9.0-10.0

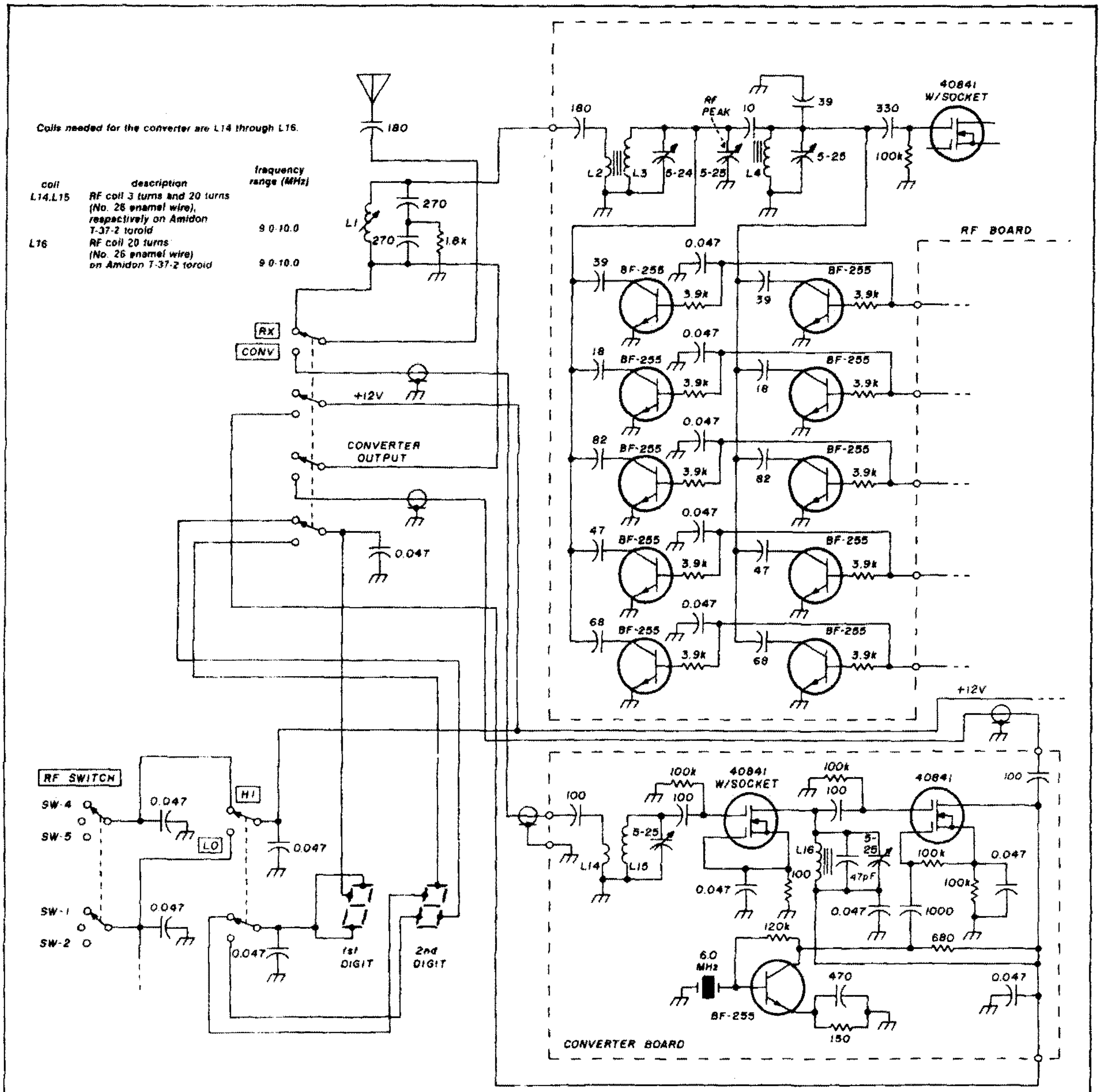

fig. 4. Schematics of the GP-78 converter and bandswitch wiring. (RF coils shown, L1 through L4, are detailed in reference 1.) Shielded bandswitch connections use miniature coaxial cable, type RG-174/U or equivalent. The 6.0 MHz crystal is a type HC-25/U by ICM. Only the first two digits (DG) of the frequency display are shown, as the third and fourth digits are wired in parallel to digit No. 2. The new control added "RF PEAK" is a 5-24 pF Jackson miniature trimmer, controllable from the front panel (see fig. 1). The remainder of this circuit (not shown) is identical to the schematics shown in reference 1. The converter shown could also be adapted to the earlier (or other) projects as a separate accessory, fitting into a 3 × 3 × 0.75 inch box, borrowing the +12 volts from the receiver.

very crowded; the addition of an extra digit would require widening the panel.

acknowledgement

Thanks go to Fernando, PY2DQU, for his support and encouragement on this project.

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ham radio

THE GUERRI REPORT

Ernie Guerri W6MGI

computer technology: fast, *fast*, *FAST!*

Some of the most dramatic changes in computer technology are taking place in architecture — that is, in the way computer logic is organized. This is necessary because electronic speeds are now so fast that the physical distance between circuit elements has become a major limitation. To alleviate this problem, the Cray supercomputer features logic bays arranged in a circle so that all interconnects are not more than one circle diameter away.

NASA has developed a design called the Massively Parallel Computer — and massive it is: over 16,000 identical processors are arranged in parallel. This approach allows an image processing task that takes 8 hours on a conventional large mainframe to be reduced to only 17 minutes! Companies such as Cray, ETA, and Fujitsu are developing computers that will be 10,000 times faster than an IBM PC by 1987 or 1988. The implications of this kind of progress make exciting news for hams. Might it someday be possible to contain a basic HF receiver on a single chip? A complete SSB receiver could actually consist of 3000 individual receivers, each having 1 kHz bandwidth and tuned to a different part of the HF spectrum under computer control. Such a unit could represent the ultimate in interference avoidance and MUF agility!

cooling high-speed circuits

Ever since the beginning of the electronic era, heat has been a problem. The absence of effective ways to remove it at the device level continues to be a major limitation to present large-scale integration. Designers are now examining methods by which an IC substrate can be bonded to a porous metal carrier, with coolant circulated through the porous metal, then evaporated and recovered in a closed system. Using this method, thermal transfer can be improved to a rate 100 times better than with conduction alone, with each LSI circuit containing its own refrigeration system.

Also significant is the interest in running VLSI circuits at the temperature of liquid nitrogen (77 degrees K). At this temperature electron mobilities go up, speed increases, and thermal efficiency improves. With modern techniques, even the cryogenic problems aren't too difficult. Look for examples of this approach to appear in commercial products before too long.

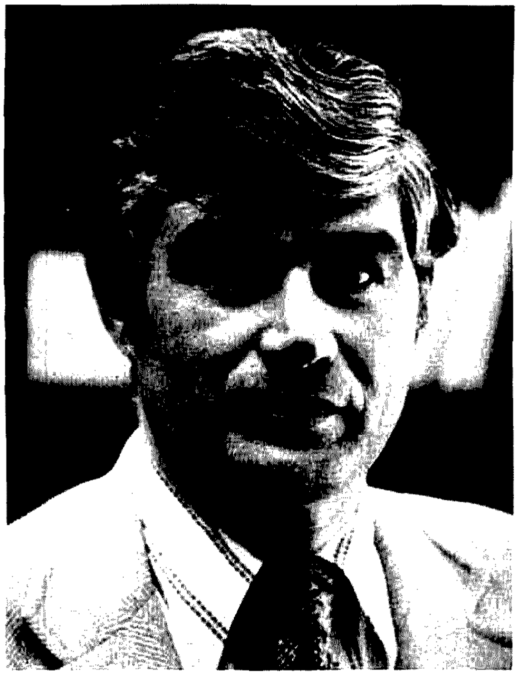
Even more exotic are super-cold devices called Josephson junctions (JJ's). These are thin film devices that operate at 4 degrees K (the temperature of liquid hydrogen), exhibit picosecond (one millionth of a microsecond) speed, and consume nearly zero power because they operate at superconducting temperatures. After spending nearly twenty years developing JJ's for supercomputer applications, IBM recently threw in the towel because of the difficulty of manufacturing the device and its support system. Work continues in Japan, however, with Fujitsu pursuing research and the Ministry of International Trade and Industry (MITI) funding the development of a JJ analog-to-digital converter. With several GHz bandwidths, such a device could make possible digital storage scopes with several hundred MHz capability, or low-noise digital VHF receivers with direct conversion to digital information at the front end. Although very low temperature devices have many attractive characteristics, they may be difficult to put to use —except in space, where the necessary low temperature is free. But the benefits are great enough to warrant a substantial continued effort around the world. We should see some exciting breakthroughs in the near future.

faster antennas, too

Take a look at what's happening in telecommunications. More and more information is being sent over each circuit; system bandwidths are being increased, and most data is now digitized before transmission. All this wideband data eventually goes to an antenna that radiates the signal. This means that the antenna has to have some measure of frequency independence — that is, be broadband. As data rates and information density increase, the pulse response of the system also becomes a consideration.

It is now being observed that many antenna types don't exhibit adequate pulse response for present and contemplated data links. The problem is not an easy one to resolve. In order to radiate, antennas must depart from the distributed characteristics that give transmission lines their good pulse response. Much attention is now being given to measuring the pulse/transient response of various antennas, and the relevant journals abound with complex math as a result. Perhaps all this effort will lead to new antennas that will couple the desired energy into space without acting as if they had all kinds of L & C hung across them.

ham radio



meet Ernie Guerri, W6MGI

Ernie Guerri, W6MGI, comes to the pages of *ham radio* with a background that includes 32 years as a licensed Amateur, and 27 years in the development and management of advanced technology. He is a Senior Member of the IEEE and a life member of ARRL.

Ernie was educated in Physics at the University of Maine, Semiconductor Electronics at the University of California, and in Business at Stanford. He has held engineering and management positions at IBM, Raytheon, and General Dynamics. Each of these positions involved work at the leading edge of aerospace technology, telementry, and deep space communications. Most recently he has been President and General Manager of the Advanced Technology Center of Gould Inc., one of America's large ($1.5B) electronics companies. In October, 1983 he left Gould to form his own technology consulting firm, which he now operates from offices in Chicago.

Ernie will be commenting on technological developments that will shape the equipment of tomorrow. Some will have direct relationship to Amateur Radio; others will hopefully encourage implementation of new technologies in yet unexplored areas of our hobby.

✓ 214

extending the modular 2-band receiver

Add two boards — and build yourself an HF transceiver for SSB and CW

In the July, 1983, issue of *ham radio* I presented a design for a practical, easy-to-use HF receiver with digital readout.[1] The simple addition of a converter to the front end extends the receiver's frequency coverage to VHF; the addition of just two more boards turns the unit into an SSB/CW transceiver as well, see **fig. 1**.

design concept

Simply stated, an SSB transmitter amplifies voice, mixes it with a carrier frequency in such a way as to balance out the carrier, removes one sideband, and mixes the result up or down in frequency to the desired output frequency.

In this design, see **fig. 2**, the audio from the microphone is amplified by a two-stage speech amplifier and applied to a simple two-diode balanced modulator that removes the carrier, providing a signal that contains two sidebands and a suppressed carrier. The carrier source is the BFO. By selecting either USB or LSB, the operating mode for the transmit signal is also selected. To remove one of the sidebands, the signal is passed through the same crystal filter used for receive. Just as the unwanted sideband is removed on receive, the output from the crystal filter contains only one sideband. Because this sideband signal, however, is too low in level to allow the transmitting mixer to function properly, an *IF* amplifier must be used to increase the signal to an effective level. The output of this stage is injected into the MC1496 double-balanced mixer IC, where it again mixes with the VFO to produce outputs at 14 and 4 MHz (9 + 5 = 14 and 9 - 5 = 4). The same filters as those used for receive are used here to remove the undesired output. The 20-meter filter removes the 80-meter signal and vice-versa.

The SSB signal present at the output of the band-pass filter is clean but at a very low level. A two-stage broadband amplifier has been designed for an output of about 10 watts (see **fig. 3**). The driver transistor is a 2N3866 which in turn drives a 2N5590 operating in class AB. The output at this power level is "clean" (low spurious/harmonic content) and requires no additional filtering. However, if you wish to drive a much higher output broadband amplifier, I would recommend adding low-pass filters for each band. Several articles on this subject have been published in this and other magazines.

operation

The same mixing scheme used for SSB transmission can be used to generate CW. A twin-T oscillator serves the dual purpose of generating both a sine wave tone, used for monitoring, and the CW signal. When a clean tone of a single frequency is applied to an SSB transmitter, a single output frequency, separated from the removed carrier frequency by the frequency of the applied tone, is produced from the transmitter. For example, if a 1 kHz tone is injected into the SSB transmitter, a CW output offset by 1 kHz is generated. Conversely, if tuning in another station produces a 1 kHz CW tone on receive, your transmitter will be on the exact transmitting frequency of the other station when you answer. (A similar method was used to produce CW in the old Heathkit SB/HW series of transceivers.)

To send CW it's necessary to activate the transmitter by either manually switching to transmit or, more easily, using the included 555 timer circuit. This keeps the transmitter on between the dots and dashes. The twin-T oscillator and the timer circuit are keyed at the same time; only the 555 timer is keyed in SSB. This timer stage switches all stages into transmit for a period determined by the adjustable time constants. In addition, the AGC for the IF amplifier must be disabled while in the transmit mode by grounding the AGC control pin 5 on the MC1350, through a 4.7 kilohm resistor.

By James J. Forkin, WA3TFS, 3210 Shadyway Drive, Pittsburgh, Pennsylvania 15227

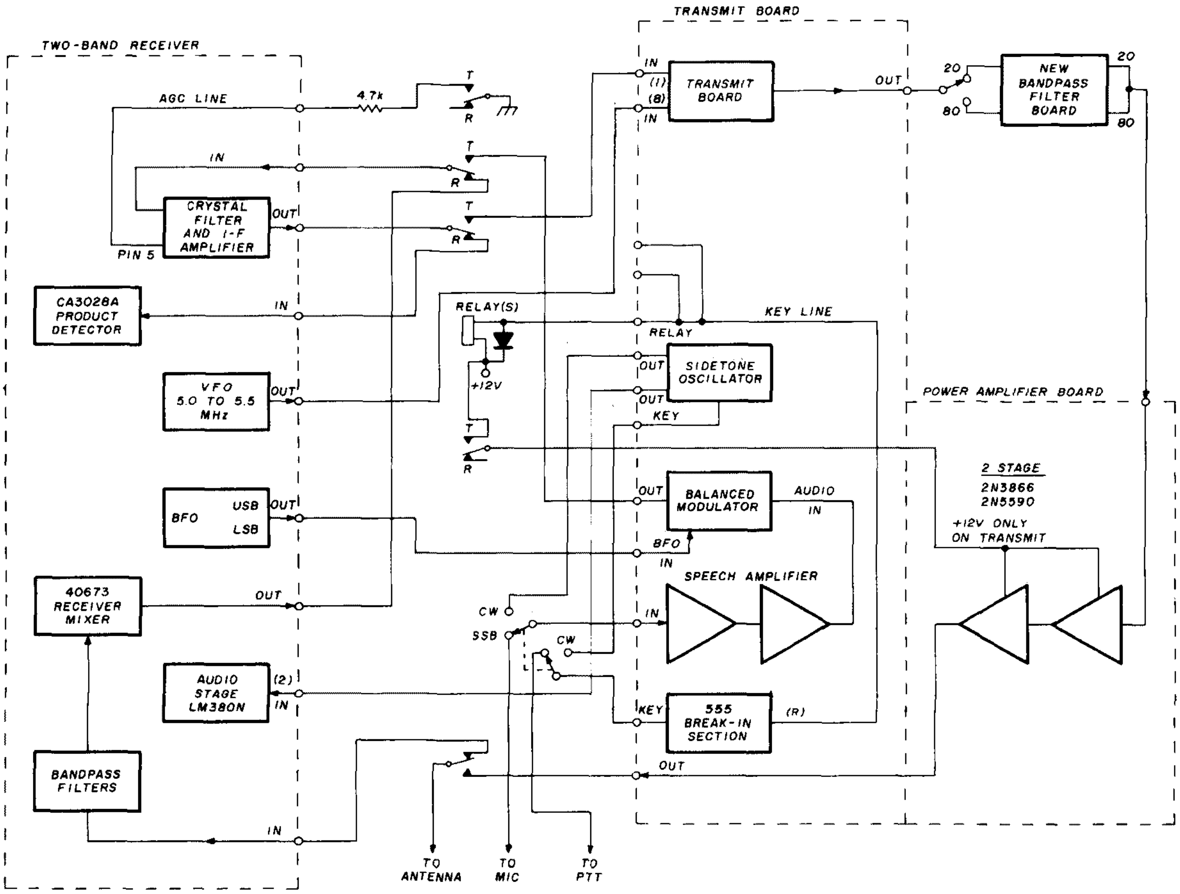


fig. 1. The interconnection diagram. Circuits within the dotted lines are on either the receive board on the left or the transmit board on the right. If more than one relay is used, wire the coil in parallel with the one shown. Use a protection diode as shown on each. Use shielded wiring on all audio circuits and 50-ohm coax on the transmit/receive relay wiring. The new bandpass filter board is identical to the board described in the July, 1983, *ham radio* receiver article.

construction

The transmit modification is accomplished through the addition of two PC boards. One consists of the two-stage amplifier described above. (Component layout and printed circuit board artwork are detailed in **figs. 4**, and **5**, respectively.) The other board, however, is the actual transmit conversion. Shown on the board, (component layout and printed circuit board artwork are detailed in **figs. 6** and **7**),from left to right, is the 1 kHz sidetone oscillator coupled through a panel switch to the speech amplifier stage in the CW mode. Next is the 555 timer stage, which holds the rig in transmit for a period of time adjusted by the trimmer pot. To the right of the timer is the transistor, used to switch the relays used in the various stages of the receiver. Next in line is the two-stage speech amplifier; note that the two stages are coupled by a jumper wire to simplify the addition of audio companders, processors, or other components later. The balanced modulator is next. Be careful to wind the toroid core exactly as shown, keeping all leads as short as possible. It is this stage that determines the ultimate level/amount of carrier suppression the transceiver will offer. The double-balanced mixer completes the board.

switching

It is important to switch the crystal filter and IF amplifier stage when going from receive to transmit. A single-pole, double-throw miniature relay mounted close to the input and output of this stage does this. Use shielded wire to and from the transmit board. (RG-174 miniature coax works well.) The front end filters must also be modified by adding two relays or alternatively, replaced with new filters exactly like those used for receive for the transmitter, thus eliminating the need for relay switching here. The only

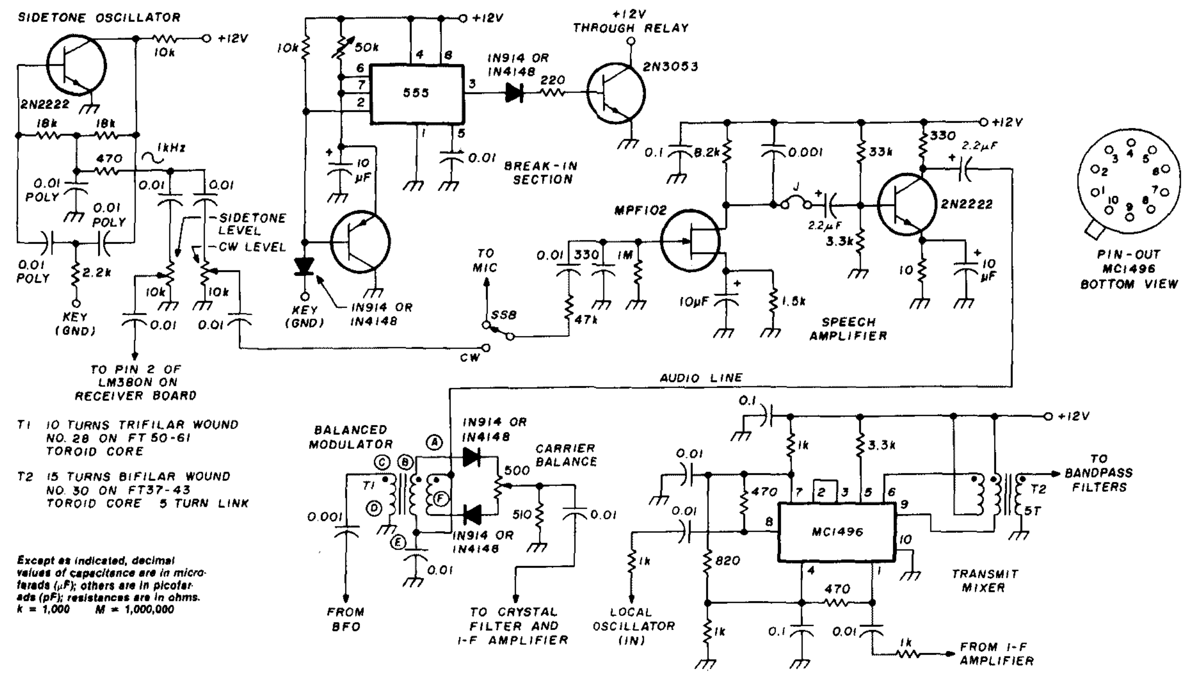


fig. 2. The schematic diagram of the transmitter board.

disadvantage to the latter approach is the need to allow for weight and additional space; despite these disadvantages, replacing the filters rather than adding relays does simplify the modification. Use coax for the filters — keep the RF where it belongs! And don't forget that to pull the AGC voltage below 5 volts for maximum gain, you'll have to use another relay or add an extra set of contacts to one of the other relays to ground pin 5 of the IF amplifier through a 4.7 kilohm resistor. Should you choose to use diode switching instead of relays, you may wish to consult several articles that have been published on the subject for help with the design.

For simplicity's sake, you may decide, as I did, to use relays. Several types of 5- and 12-volt DC relays are available on the surplus market. Use 12-volt relays if you can find them at a reasonable price, or wire three 5-volt relays in series; they'll key reliably on 12 volts. Mount your relays to the board with double-sided foam tape or glue. Place a diode across each relay coil to prevent voltage spikes.

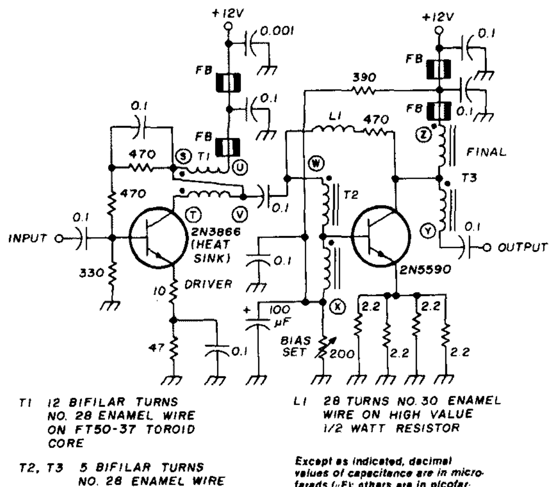


fig. 3. The schematic diagram for the final amplifier board. The driver section is similar to the VFO buffer in the receiver and the final stage is patterned after the original design.

initial adjustments

After completing the modification, make sure that the receiver still works. Realign it and check the BFO frequencies. When you're convinced that the receiver is working as well as it did before the modification, connect a dummy load to the antenna terminal and key the transmitter in the CW position. Check that all relays switch as they should and adjust the 555 for a hold-in time of about 1 second. Check that the AGC voltage at pin 5 of the MC1350 is in fact dropping below 5 volts on transmit. Put the rig into SSB and

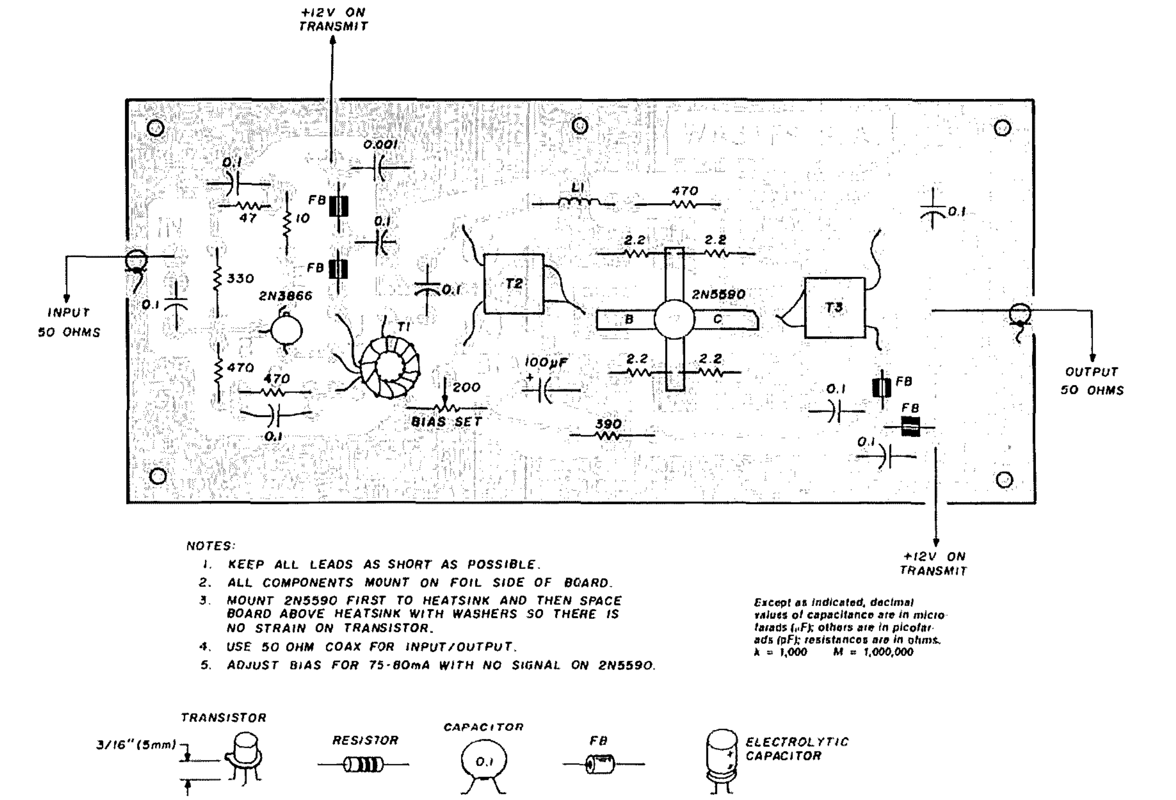


fig. 4. The parts layout for the RF amplifier board. Positive 12 volts is applied only on transmit. Set bias adjustment for an idle current in the final stage of 75 to 80 mA with no drive applied to the board.

adjust the output stage for a resting current of 80 mA. With the rig in SSB there will be no drive to the final stage. Put the switch back in CW. When keying, adjust the trim pot in the oscillator stage for sidetone level. Increase the drive level with the other trim pot until no increases in output level from the transmitter are noted. Back off the adjustment slightly. CW tune-up is completed.

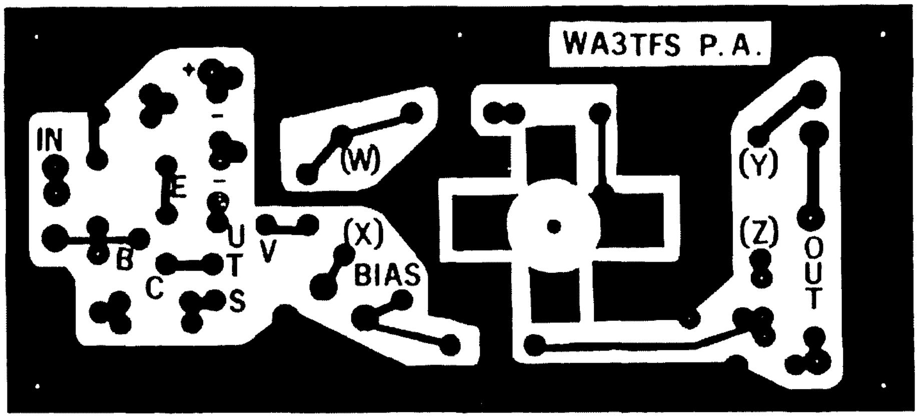


fig. 5. The foil side layout for the RF amplifier board. All parts on this board are mounted on the foil side.

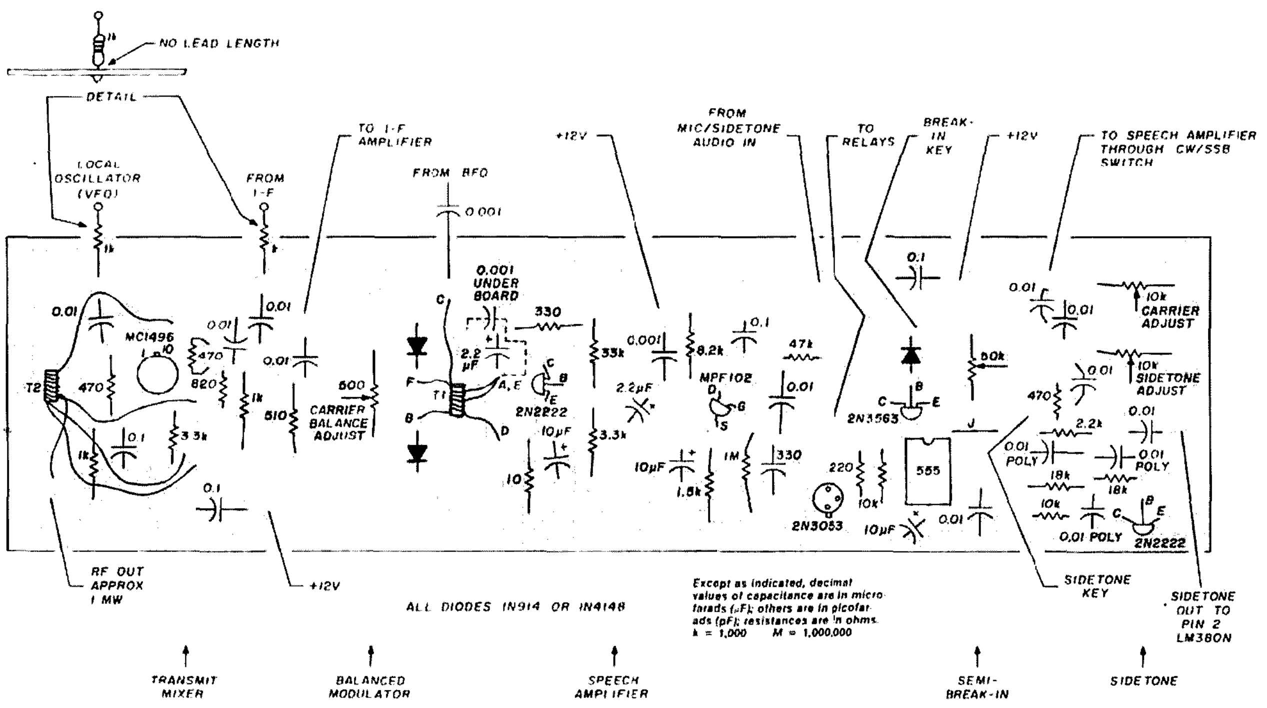


fig. 6. The parts layout for the transmit board. (Note: On this board components are *not* mounted on the foil side.) Use shielded wire for all audio connections, and 50-ohm coax for all RF wiring.

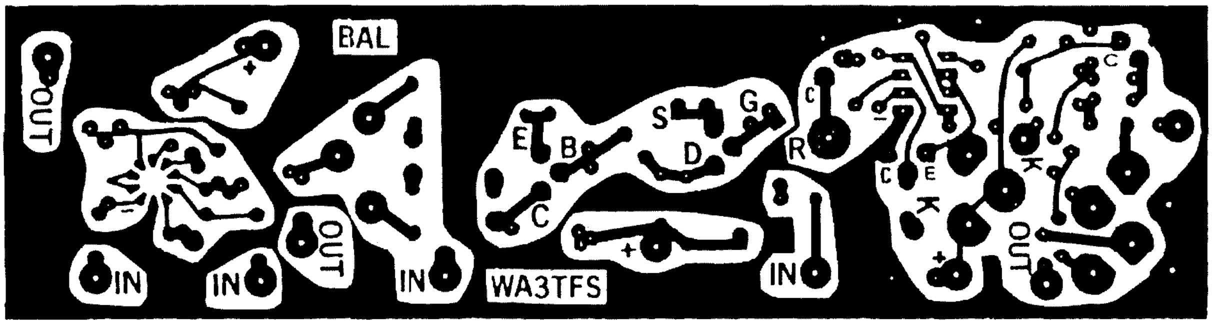


fig. 7. The foil side layout for the transmitter board.

Next attach a microphone to the input. (A CB replacement microphone will be sufficient.) If the CW/SSB switch has been wired correctly, the microphone will be connected to the input of the speech amplifier and the twin-T oscillator will be disconnected. When you press the PTT and whistle into the microphone, the RF output should increase. The level should be about the same as when transmitting CW, but may vary because of different output levels of various microphone elements.

Disconnect the microphone and while still in SSB, key the transmitter with a jumper wire. While checking the output with a meter, or better still, an oscilloscope, adjust the trim pot in the balanced mixer for minimum output and, consequently, maximum carrier suppression. If you can't see any change in the meter reading as you make this adjustment, you'll know you've either wound the coil in the balanced modulator incorrectly or caused some stage to oscillate because of poor wiring layout or failure to ground something. Check your construction step by step. You should see a definite null in output power. If everything

appears to be in order, go back and readjust everything until no further change occurs. Your new transceiver is now ready to be connected to the antenna.

conclusion

This complete transceiver will operate reliably and efficiently as long as care is taken to attach a matched 50-ohm load. The output stage will not self-destruct if you have high SWR or forget to attach the antenna, but output power will be low. The rig should run about 8 to 10 watts out into a matched antenna. I have worked all states on 20-meter sideband and find I require no more power from the home station.

A kit is available from the author to make the modification described in this article. For details, please send an SASE to me at the address shown at the beginning of this article.

ham radio

receiving signals from space

How to locate geostationary satellites from your QTH

With the **price of TV-Receive Only (TVRO)** terminals on its way down and the availability of channels expanding, interest in geostationary satellites is increasing. This article describes how to locate these satellites from any given latitude and longitude in terms of azimuth, elevation, and range.

Two programs are included — one in BASIC for the TRS-80™ (level II or similar) and the other for the Hewlett-Packard 67 or equivalent. While the mathematics are the same for each program, some minor changes have been made to accommodate the specific programming language used and the functions available on each machine. For example, (cos/sin) is substituted for (cot) because of the absence of the cotangent function on the HP-67 and in TRS-80 Level II BASIC.

celestial mechanics

For a satellite to always appear stationary above a particular point on earth, it must have the same period as the Earth — that is, 23 hours, 56 minutes, 4.09 seconds or 86164.09 seconds. In order to have a period that matches that of the earth, the geostationary satellite must be a specific height above earth. This measurement can be found by using the Newtonian law *stating that the square of the* velocity of an object (satellite in this case) is equal to the universal gravitational constant times the mass attracting the object (the Earth), divided by the distance of the object from the center of the mass (Earth).

$$V^2 = \frac{GM_E}{D} \tag{1}$$

where V = velocity of satellite
G = universal gravitational constant
$6.673 \times 10^{-11} \frac{\text{Newtons-meter}^2}{\text{kilogram}^2}$
M_E = Mass of Earth = 5.975×10^{24} kg
D = distance from center of earth to satellite
= (R + H) = radius of earth + height of satellite above earth

G and M_E are constants and can be combined:

$$G' = GM_E = 3.987 \times 10^{14}\ meters^3/sec^2$$
$$= 3.987 \times 10^5\ km^3/sec^2 \tag{2}$$

which results in

$$V^2 = \frac{G'}{R+H} = \frac{G'}{D} \tag{3}$$

$$\text{or } V_{SAT} = \sqrt{\frac{G'}{D}} \tag{4}$$

The period of one complete revolution of the satellite is equal to the distance it travels in orbit divided by its velocity or:

$$P_{SAT} = \frac{2\pi D}{V_{SAT}} = \frac{2\pi D}{\sqrt{\frac{G'}{D}}} = 2\pi\sqrt{D^3/G'} \tag{5}$$

But this is equal to 86,164 seconds (approximately 24 hours) for it to be a *geostationary* satellite as explained above.

By Dennis Mitchell, K8UR, 1 Cider Mill Lane, Upton, Massachusetts 01568

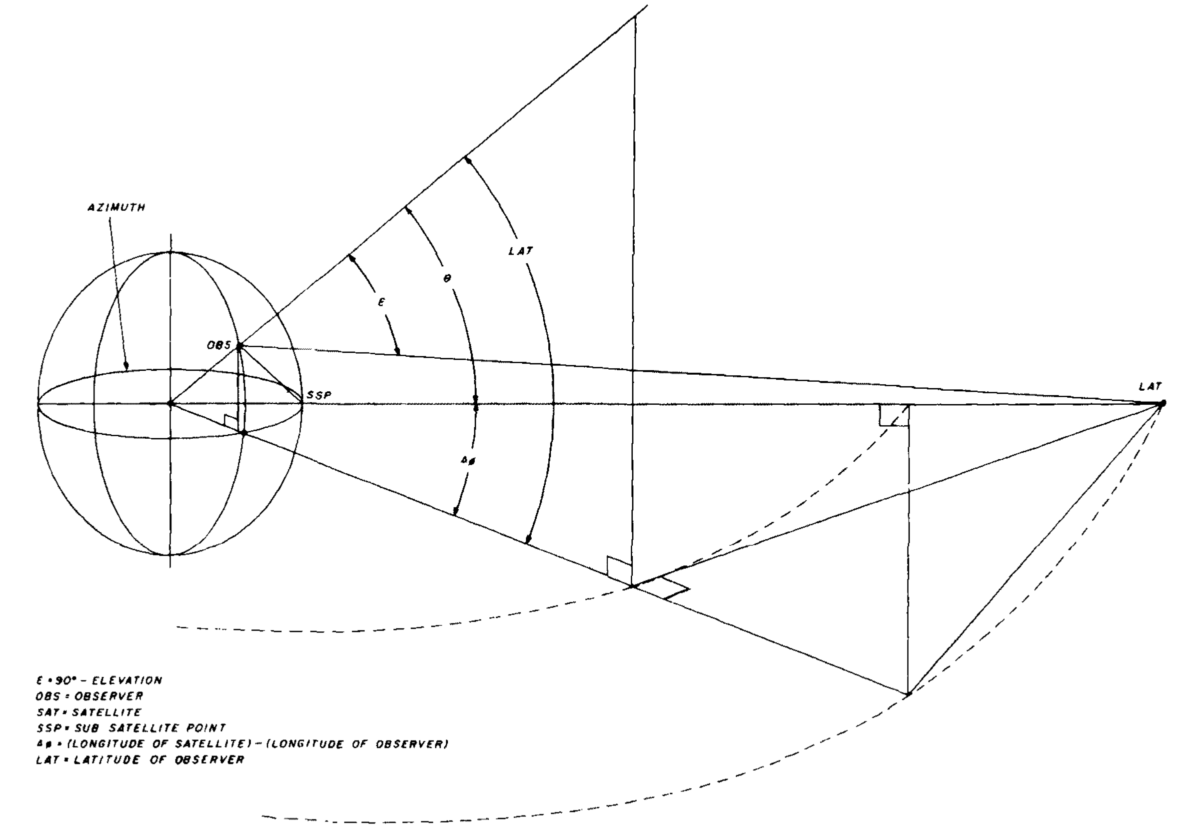


fig. 1. Angles involved in calculating satellite's azimuth, elevation and range.

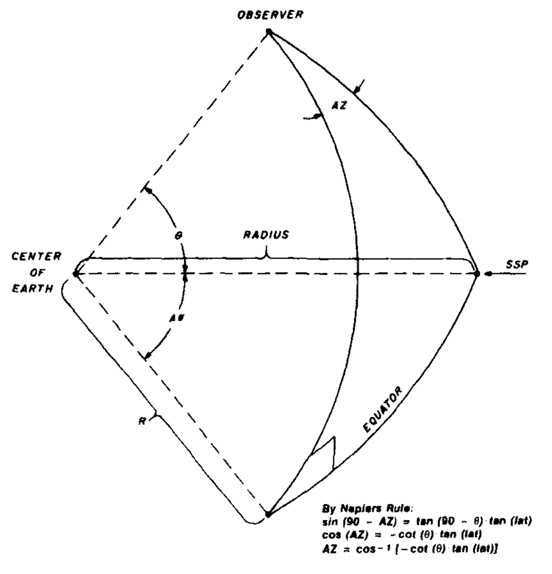


fig. 2. Napier's rule illustrates the relationship between the satellite and the observer.

Rearranging terms and solving for D:

$$D = \sqrt[3]{\left(\frac{P_{SAT}}{2\pi}\right)^2 G'} \qquad \textbf{(6)}$$

$$= 42{,}168 \ km$$

$$H \ (\text{height of satellite above earth}) = 42{,}168 - 6378$$
$$= 35{,}790 \ km$$
$$= 22{,}239 \ miles$$

$$\text{and } V_{SAT} = \sqrt{\frac{G'}{D}} = \frac{398{,}700}{42{,}168} = 3.075 \ km/sec$$
$$= 3075 \ meters/sec.$$

finding azimuth, elevation, and range

Fig. 1 shows the angles involved in finding the azimuth, elevation, and range of the satellite.

Fig. 2, an exploded view of a section of the Earth, shows how the locations of the Earth's center, an observer, the Equator, and latitude are related. The difference in longitude and the sub-satellite point are also shown.

By viewing **figs. 1** and **2** and using Napier's rule with

table 1. List of the current C-band (3.7-4.2 GHz) geosynchronous satellites and their longitude.

satellite name	degrees west
AURORA I	143
ANIK B	109
ANIK C2	105
ANIK D	104.5
ANIK III	114
COMSTAR I	128
COMSTAR II	95
COMSTAR III	87
COMSTAR IV	127
GALAXY I	128
GALAXY II	74
SATCOM I-R	139
SATCOM II-R	72
SATCOM III-R	131
SATCOM II	119
SATCOM IV	83
SBS I	100
SBS II	97
SBS III	95
TELSTAR 301	96
USAT I	85
WESTAR 1	99
WESTAR II	79
WESTAR III	91
WESTAR IV	99
WESTAR V	123

table 2. Sample calculation of geosynchronous satellite azimuth, elevation, and range for an observer in Upton, Massachusetts.

satellite	longitude	azimuth	elevation	range (km)
AURORA I	143	257	5	41128
ANIK B	109	228	28	38763
ANIK C2	105	224	30	38555
ANIK D	104.5	223	30	38530
ANIK III	114	233	25	39049
COMSTAR I	128	245	15	39979
COMSTAR II	95	212	35	38125
COMSTAR III	87	202	38	37883
COMSTAR IV	127	244	16	39907
GALAXY I	128	245	15	39979
GALAXY II	74	183	41	37700
SATCOM I-R	139	254	7	40811
SATCOM II-R	72	180	41	37695
SATCOM III-R	131	248	13	40199
SATCOM IV	83	196	39	37799
SBS I	100	218	33	38324
SBS II	97	215	34	38201
SBS III	95	212	35	38125
TELSTAR 301	96	213	35	38162
USAT I	85	199	39	37838
WESTAR II	79	190	40	37739
WESTAR III	91	207	37	37993
WESTAR IV	99	217	33	38281
WESTAR V	123	241	19	39627

```
10 CLEAR 100
20 REM*********************************************************
30 REM***  THIS PROGRAM WILL COMPUTE THE AZIMUTH, ELEVATION ***
40 REM***  AND RANGE TO GEOSTATIONARY SATELLITES FOR TVRO   ***
50 REM********* BY D.C.MITCHELL-K8UR/1 1/1/81 ****************
52 REM********* UPDATED SATELLITE LIST 6/27/84 ***************
55 REM******** 1 CIDER MILL LN. UPTON,MA.01568 ***************
60 REM*********************************************************
70 CLS
80 PRINTCHR$(23):PRINT"ENTER YOUR LATITUDE":GOSUB 100:LA=DD
90 CLS:PRINTCHR$(23):PRINT"ENTER YOUR LONGITUDE":GOSUB 100:LO=DD:GOTO 150
100 PRINT@260,"DEGREES-----":PRINT@324,"MINUTES-----":PRINT@388,"SECONDS-----"
110 PRINT@286,"";:INPUTDD:PRINT@324,"MINUTES----- ";:INPUTMM:PRINT@388,"SECONDS----- ";:INPUTSS
120 PI=3.14159
130 LA$=STR$(DD)+STR$(MM)+STR$(SS):DD=(MM*60+SS)/3600+DD:RETURN
140 REM
150 INPUT"ENTER YOUR CITY";C$:INPUT"ENTER YOUR STATE";T$:C$=C$+","+T$
160 CLS
170 LA=PI*LA/180:LO=PI*LO/180
180 PRINT"    SATELLITE ***  AZIMUTH / ELEVATION / RANGE *** LOCATOR"
190 PRINT"                     FOR   "C$:PRINT""
200 PRINT"SATELLITE    LONG       AZIMUTH        ELEVATION     RANGE (KM)"
210 PRINTSTRING$(60,"*")
220 READ S$,S
230 IF S$="END" THEN END
240 S=S*PI/180
250 DEFFN AC=-ATN(X/SQR(-X*X+1))+1.5708
260 X=(COS(LA)*COS(S-LO)):TH=FNAC
270 X=(-TAN(LA)*COS(TH)/SIN(TH)):AZ=FNAC
280 IF SIN(S-LO)>0 THEN AZ=6.28-AZ
290 EL=ATN((COS(TH)-.151046)/SIN(TH))
300 RA=SQR((1.81854E9-(5.37111E8*COS(TH))))
310 AZ=AZ*180/PI:EL=EL*180/PI:S=S*180/PI
320 DATA"AURORA I",143,"ANIK B",109,"ANIK C2",105,"ANIK D",104.5
321 DATA"ANIK III",114,"COMSTAR I",128,"COMSTAR II",95,"COMSTAR III",87
322 DATA"COMSTAR IV",127,"GALAXY I",128,"GALAXY II",74,"SATCOM I-R",139
323 DATA"SATCOM II-R",72,"SATCOM III-R",131,"SATCOM IV",83,"SBS I",100
324 DATA"SBS II",97,"SBS III",95,"TELSTAR 301",96,"USAT I",85,"WESTAR IV",99
325 DATA"WESTAR II",79,"WESTAR III",91,"WESTAR V",123,"END",999
326 AZ=FIX(AZ):EL=FIX(EL):RA=FIX(RA)
330 PRINTS$;TAB(13)S;TAB(24)AZ;TAB(39)EL;TAB(53)RA
340 GOTO 220
```

fig. 3. BASIC language program listing used to determine geosynchronous satellite azimuth, elevation angle, and range from your QTH.

spherical trigonometry and some trigonometric identities, the equations for azimuth, elevation, and range are:

$$\theta = cos^{-1}[cos\ (lat) \cdot cos(\Delta\ long)] \quad (7)$$

$$Az = cos^{-1}\ [-tan(lat)\ cos\ (\theta)/sin\ (\theta)] \quad (8)$$

If sin (Δlong) > 0 then $Az = 360 - Az$ and the

$$elevation\ angle = tan^{-1}[cos(\theta) - 0.151)/sin(\theta)] \quad (9)$$

where: $R/(R + H) = 6378/42168 = 0.151$
and:

$$Range = \sqrt{(R + H)^2 + R^2 - 2 \cdot (R + H) \cdot R\ cos\ \theta}$$

program hints

In the HP-67 program, the observer's latitude and longitude are replaced in decimal form. (Latitude is replaced in lines 3 through 7; longitude in lines 9 through 13.) Don't forget to use your own numbers — not mine — in these steps. The only other entry is the satellite longitude taken from **table 1**; after entry, hit key (A). Outputs are elevation, azimuth, and range in that order.

The BASIC program, which should need no explanation, prompts the user for all inputs. As shown in **table 2**, outputs provide satellite name, azimuth, elevation, and range in kilometers.

Locating the geostationary satellite you're looking for among the many orbiting the Earth in the crowded "satellite belt" is getting more difficult, but with a computer program such as this and some good microwave gear, they *can* be found.

STEP	KEY ENTRY	KEY CODE	COMMENTS
001	f LBL A	31 25 11	
	STO A	33 11	
	4	04	
	2	02	USERS OWN
	.	83	LATITUDE
	4	04	
	8	08	
	STO 0	33 00	LATITUDE
	7	07	
010	1	01	USERS OWN
	.	83	
	2	02	LONGITUDE
	2	02	
	STO 1	33 01	LONGITUDE
	RCL A	34 11	
	RCL 1	34 01	
	-	51	
	STO B	33 12	Δ LONGITUDE
	f COS	31 63	
020	RCL 0	34 00	
	f COS	31 63	
	x	71	
	g COS^{-1}	32 63	
	STO 2	33 02	θ
	f COS	31 63	
	RCL 0	34 00	
	f TAN	31 64	
	CHS	42	
	x	71	
030	RCL 2	34 02	
	f SIN	31 62	
	÷	81	
	g COS^{-1}	32 63	
	STO C	33 13	AZIMUTH
	f x > 0	31 81	
	GTO 0	22 00	
	f LBL B	31 25 12	
	RCL 2	34 02	
	f COS	31 63	
040	.	83	
	1	01	
	5	05	
	-	51	
	RCL 2	34 02	
	f SIN	31 62	
	÷	81	
	g TAN^{-1}	32 64	
	STO D	33 14	ELEVATION
	f - x -	31 84	
050	RCL C	34 13	
	f - x -	31 84	
	GTO 1	22 01	
	f LBL 0	31 25 00	
	3	03	
	6	06	
	0	00	
	h x <> y	35 52	
	-	51	
	STO C	33 13	
060	GTO B	22 12	
	f LBL 1	31 25 01	
	1	01	
	.	83	
	8	08	
	1	01	
	8	08	
	EEX	43	
	9	09	
	ENTER	41	
070	5	05	
	3	03	
	7	07	
	.	83	
	1	01	
	4	04	
	1	01	
	EEX	43	
	6	06	
	RCL 2	34 02	
080	f COS	31 63	
	x	71	
	-	51	
	f √	31 54	
	R/S	84	RANGE
090			
100			
110			

REGISTERS

0	1	2	3	4	5	6	7	8	9
LAT	LONG	θ							
S0	S1	S2	S3	S4	S5	S6	S7	S8	S9

A	B	C	D	E	I
SAT LONG.	Δ LONG.	AZIMUTH	ELEVATION		

fig. 4. HP-67 program listing for locating geosynchronous satellites.

ham radio

ham radio TECHNIQUES

Bill Orr W6SAI

more about radials

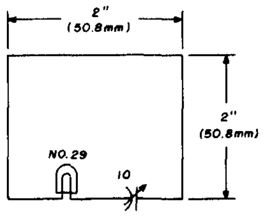
Antenna experimentation is one field in which the enthusiast doesn't need an advanced degree in electronics and a room full of expensive test equipment. Sometimes a twenty-five cent "instrument" can provide meaningful results for the investigator.

A case in point: the experiments of Ralph, W8HXC, and Don, AF8B, which were designed to determine the effectiveness of quarter-wave radials on various 2-meter vertical antennas. The tests, conducted intermittently over a period of 5 years, pointed out some interesting aspects of radials that help to remove some of the mystery of VHF antennas.

The original investigation was designed to determine the best way to decouple the shield of a coaxial feedline from the field of the VHF antenna. The goal was to make the antenna do all the work, and to prevent the feedine from becoming part of the antenna. Only by making the feedline "inert" to the field of the antenna could the antenna do its job of laying down a low-angle signal.

To determine the degree of RF on the outer surface of the coaxial line, the simple "RF-sniffer" shown in **fig. 1** was built. It was used to detect current loops on the antenna elements, the feedline, and supporting mast and structure. Made out of junk box parts, the simple device worked well with transmitter powers as low as 7 watts.

The "sniffer" consisted of a 144-MHz resonant circuit with a pilot lamp indicator, all mounted on a wooden handle. The capacitor was adjusted for maximum glow of the lamp (resonance) when held near the RF source used in the experiments.



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

fig. 1. The RF "Sniffer" for 144 MHz is built on a cut-down ping-pong paddle. Capacitor is adjusted for greatest lamp illumination at the test frequency.

The first experiments conducted were on a homebrew 1/4-wave groundplane antenna. It was found that the outside of the coax line, which dropped down beneath the groundplane antenna, was "hot" and exhibited a standing wave of energy along it that could be detected with the "sniffer." Excellent feedline isolation was achieved by simply wrapping the RG-58/U feedline into a two-turn coil 1-1/2 inches in diameter directly below the antenna. This little RF choke decoupled the feedline so that it was isolated from the antenna.

The next experiment was with an extended *half-wave vertical antenna*. RF was found on the feedline, and adding the choke in the feedline accomplished little. The outside of the line was still coupled to the antenna. Four quarter-wave radial rods were added to the antenna immediately below the matching coil (**fig. 2**). It was necessary to readjust the antenna for best SWR; however, the feedline isolation was *not* improved, and the radials did not seem "hot"with RF energy.

The last experiment, which was more meaningful, used a 5/8-wavelength antenna (48-inch long radiator) and a two-turn base matching coil (**fig. 3**). The feedline was carried down inside the metal supporting mast and a set of four quarter-wave radials with a clamping arrangement that allowed the radials to be placed anywhere on the mast was added.

Initially, the radials were positioned at the base of the antenna loading coil and the coil feedpoint was adjusted for best SWR indication. When a near-perfect match was achieved, the RF "sniffer" was used to examine the feedline. Unhappily, the feedline and mast indicated pronounced current loops over the entire length! The feedline and support pipe had become part of the antenna in spite of the radials, which were supposed to isolate the antenna from the feedline. In addition it was discovered that there was very little RF in the radials, a

sure indication they were not doing their job.

Further experimentation proved that moving the radials down the mast, *away* from the antenna base, changed the SWR reading and required feedpoint readjustment. By cut-and-try a combination of feedpoint adjustment and radial position yielded excellent SWR, radials "hot" with RF and no detectable current loops on either the feedline or the supporting mast below the radials. Measurement placed this optimum radial position 3/8-wavelength below the base of the antenna. The radial angle was finally set at 45 degrees to the horizontal for best SWR.

Further tests with this antenna and with a car-mounted antenna of the same general type led to the interesting discovery that 5/8-wavelength long radials attached at the base of the 5/8-wavelength antenna provided the same excellent feedline isolation as did 1/4-wavelength radials attached 3/8-wavelength down the structure. A final experiment showed that radials could be attached to the mast at any point up to 3/8-wavelength beneath the antenna base provided that the sum of radial length and distance from the antenna base totalled 5/8-wavelength.

Don, AF8B, points out that the 5/8-wavelength vertical antenna plus the 5/8-wavelength long radial system is the same *overall electrical length as* an extended double-Zepp antenna.

The conclusion of the experiments is that radial length cannot be taken for granted and, in the case of an extended antenna, may not be 1/4-wavelength long. The important dimension is the overall length of antenna plus radial. The test to determine radial length is to use a "Sniffer" to make sure the RF remains in the radials and not on the outside of the coaxial feedline. (Thanks to Don, AF8B, for supplying data on the W8HXC and AF8B experiments.)

the Australian wideband dipole

Reader interest has been aroused by my description of the so-called "Australian dipole" wideband antenna (January, 1983, page 67). It seems that there is a whole family of wideband HF antennas and other related products of this type manufactured by *Antenna Engineering Australia PTY. Ltd.*, Box 191, Croydon, Victoria 3136, Australia. Contact Ian R.H. Wade, Sales Manager, for further information. The correct name of the antenna described in my January column is *Model 632 Travelling Wave Dipole*.

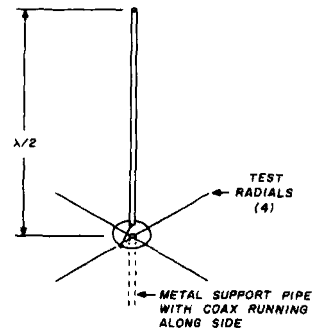


fig. 2. Test radials were added to extended 1/2 wave antenna.

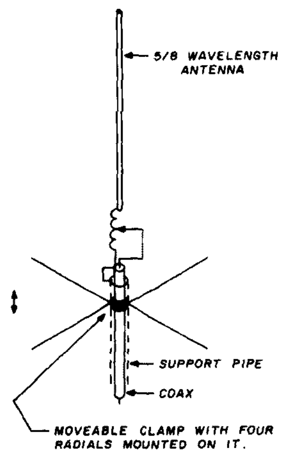


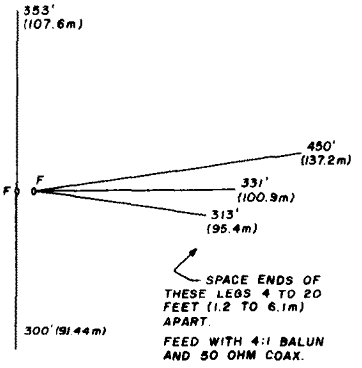
fig. 3. 5/8-wavelength "gain" antenna was mounted to support pipe and feedline passed down inside the pipe. A set of 1/4-wave radials with a mounting clamp was placed on the pipe. Radials could be moved up and down with reference to the base of the antenna.

the K4EF "all-band" antenna

Several years ago Ev Brown, K4EF, described a wire antenna that would cover all HF Amateur bands between 80 and 10 meters (*ham radio*, May 1977, page 10). Since then he's done a lot of work on his design and has devised a new configuration that has several advantages over the old one. The new antenna covers the 160-meter band, uses four support points instead of five, and occupies less space. In addition, because the elements are arranged in a V-configuration, it provides some signal gain on the higher frequency bands.

A plan view of the new antenna design is shown in **fig. 4**. The array consists of five long wires arranged in a semicircle. The antenna is fed at points F-F with a 4-to-1 balun and a 50-ohm transmission line. In actual use, one of the two elements at the left of the illustration is used with one of the three wires at the right. The wires can be selected from the operating position with a remote switch. For example, if the 353-foot wire is added to the 313-foot wire, an element 666 feet long is produced. An odd number of half waves is required to produce approximately 200 ohms feedpoint impedance at or near the element center. The chart of **table 1** shows the odd-halfwave resonances in this combination. As can be seen, the bandwidth coverage is enormous (see column 3), and when you consider that the 666-foot combination is merely one of six, the complete configuration provides wide spectrum coverage with very low SWR. A simple computer program could calculate all of the resonances and bandwidths for all elements. The results could then be combined to determine what frequency gaps (if any) exist in the complete array coverage.

As Ev says, ". . . it is difficult to convey to a ham who has never used an all-band, broadband antenna just how convenient it is. During contests, changing bands is accomplished by flipping the bandswitch. Checking band conditions is done in an instant.



element length feet	Amateur band resonant frequencies 160	80	40	30	20	18	15	12	10
613		3970	7180	18.4		18.4		24.8	28.0
831		3950	6950	10.1		17.9	21.0	24.1	28.8
666		3050			14.0		21.4		30.2
694		3585							29.4
750	1940								
903	1820								

fig. 4. Two legs of antenna comprise an element. For example, the 353-foot leg plus the 313-foot leg form an element 666-feet long. The chart shows resonance at 3650 kHz, 14.0 MHz, 21.4 MHz, and 30.2 MHz. Other combinations provide additional resonant frequencies. (Top view of antenna shown.)

My FOC friends frequently ask to get credit for another band and find me waiting for them. Perhaps the most important aspect of the idea is that it encourages the operator to use the whole spectrum available."

the W2TBZ quad-loop beam antenna

I had not seen Sid, W2TBZ, for over 15 years and our QSOs on the air were few and far between. "Keep In touch," I had said, and just recently I heard from him — with a new antenna idea that he was using with great success on 15 and 20 meters.

To stay in touch with his friends, Sid needed an inexpensive wire beam that could be easily erected and would provide a modest amount of gain and a low angle of radiation. Various antennas were tried, and the final version, a 2-loop Quad beam is shown in fig. 5. Estimated gain of this bidirectional array is about 4.5 dB over a dipole.

The antenna consists of two side-by-side Quad loops, horizontally polarized and driven in phase. The feed system consists of two equal lengths of 300-ohm TV line and a 1-to-1 balun. The feedpoint impedance of a single loop in this configuration runs about 120 ohms, so parallel-connected loops provide a terminal impedance close to 60 ohms. This provides a good match to a 50-ohm transmission line system.

The 40-foot masts support the antenna. The figure-8 radiation pattern is at right angles to the plane of the array. The pattern is sharper than that of a dipole, being about 60 degrees between the half-power (−3 dB) points.

table 1. Odd halfwave resonances in 666 feet of wire.

band meters	electrical length halfwaves	resonant frequency (MHz)	bandwidth to 2:1 SWR points
80	3	2.179	2.142 to 2.216
	5	3.657	3.597 to 3.717
	7	5.134	5.057 to 5.211
	9	6.612	6.512 to 6.712
	11	8.089	7.969 to 8.209
	13	9.567	9.427 to 9.707
	15	11.044	10.879 to 11.184
	17	12.522	12.342 to 12.702
20	19	13.999	13.789 to 14.209
	21	15.477	15.244 to 15.709
	23	16.954	16.999 to 17.208
	25	18.432	18.155 to 18.709
	27	19.909	19.610 to 20.207
15	29	21.386	21.065 to 21.706
	31	22.864	22.521 to 23.206
	33	24.341	23.975 to 24.706
	35	25.819	25.431 to 26.206
	37	27.296	26.886 to 27.705
10	39	28.774	28.342 to 29.205
	41	30.251	29.797 to 30.704

Note: The 666-foot element (summarized above) is only one of six element combinations. Single element switch will provide enormous coverage of HF spectrum with low SWR.

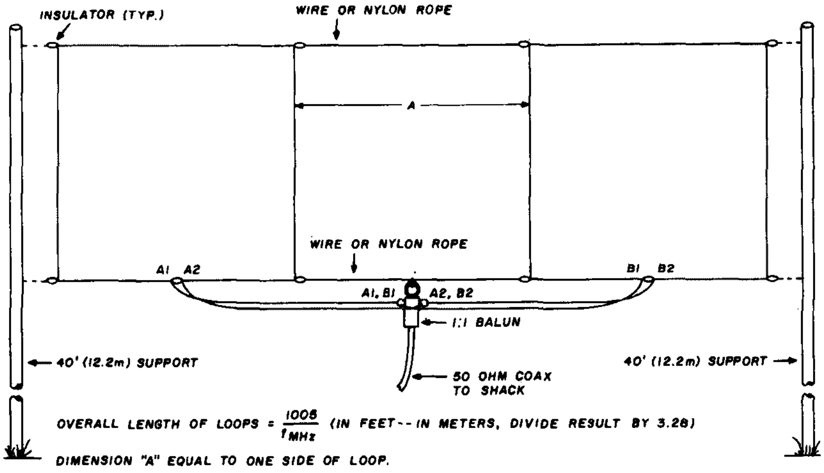


fig. 5. The W2TBZ loop beam. The phasing lines are made of 300-ohm TV line. Length of phasing line = $\frac{468}{f_{MHz}}$ × velocity factor of 300-ohm line. Balun is supported at center of lower nylon rope.

interesting reading!

From time to time I like to recommend interesting books or periodicals that provide information that otherwise may be unobtainable, and that are of general interest to Radio Amateurs.

This month's recommendation is *The Monitoring Times*, published monthly by Grove Enterprises, Inc., 140 Dog Branch Road, Brasstown, North Carolina 28902. The subscription rate is $10.50 for one year.

The Monitoring Times is full of timely information about what's going on in the HF/VHF spectrum. The editor and publisher is Bob Grove, WA4PYQ. This newspaper covers items of interest not generally found in Amateur publications. I look forward with interest to each issue! The latest information on the mysterious "beacon" and "numbers" stations may be found in this publication, as well as up-to-date information and interesting stories of other aspects of radio communication.

Some of the columns in *Monitoring Times* are "High Seas Radio," "Signals from Space," "Utility Intrigue," "RTTY/FAX," and "Pirate Radio." There's also a good review of some of the new communications receivers in the present issue of this interesting publication.

Good luck, Bob — you have a winner!

ham radio

make your own PC boards
using silk screen techniques

A step-by-step guide to easy, inexpensive duplication of simple circuits

At least two silk-screen approaches to PC board duplication will work for reproducing relatively simple circuits. One is based on the use of printing film, and the other is based on the use of a photo-sensitizing material that can be applied directly to a silk screen or to a board. I have found both of these techniques to be quite satisfactory, and I consider them to be superior to the usual board photo-sensitizing approach for all but the most sophisticated circuit configurations. The screen-sensitization technique can be used with "LIFT-IT"™ patterns or by applying sensitizer directly on the board if one is certain replication will not be necessary. The printing-film approach is suitable for relatively simple circuits such as those used for RF voltmeter probes.

In order to produce a conductive pattern on a circuit board it is necessary to transfer a drawn pattern to the copper on a board. This requires the application of material that will protect the desired conductor area from an etchant. Of the various methods available, silk-screen techniques are probably the least expensive and most convenient solution to the typical multi-board problems encountered by hams. (Where only single boards are required, the photo-sensitizing method can be applied directly to the board.)

board preparation

I buy my copper-clad material (copper one side) at hamfests, usually for less than one cent per square inch, far less than the 20 cents or more charged for sensitized boards.

First I cut the board to size with a bandsaw. Metal shears or a pair of tinsnips can also be used. The board may also be scored with a linoleum knife and separated. After the boards have been cut to size, the edges and corners should be deburred to avoid cutting the silk in the process of inking. If the board is badly corroded the copper surface should be scoured with 600 grit emery, and finally with a cleaning powder such as "Old Dutch Cleanser," one that is free of chlorides and phosphates. An all-over clean copper lustre is required to assure efficient etching.

mounting frame preparation

Two kinds of mounting frames can be used. Because the boards I use are seldom larger than 3 by 5 inches, I purchased some 3/4-inch square wood strips and cut them into 6- and 8-inch lengths. Using picture frame clamps, I assembled these pieces into frames having outer dimensions of about 7 by 9 inches, gluing the pieces of wood together with white glue and inserting 2-inch long wood screws through the joints and reinforcing the joints with flat L brackets measuring 1-1/2 inches on each side (see **fig. 1**). After assembly, the forms should be protected with shellac to improve their resistance to water. I use these frames for board applications having continuous use, such as power supply configurations.

It is difficult to get enough tension on the screen to minimize under-flow with this arrangement. I have found it convenient to attach 7-inch pieces of flat aluminum stock about 3/4 of an inch wide on the inside of the long sides. These can be used to stretch the screen tightly. Beware of sharp corners on the tensioning bars; any burrs or sharp edges or corners will cut the silk. I cut slots in the bars and use screws to hold them in place. Much less underflow results.

By Keats A. Pullen, Jr., W3QOM, 2807 Jerusalem Road, Kingsville, Maryland 21087

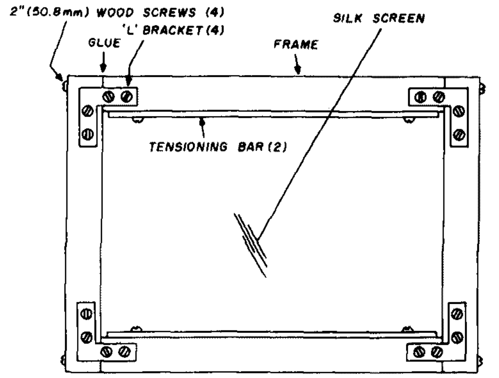


fig. 1. Wood silk-screen printing frame with pattern. The irregular outer edge of the wood frame is caused by the silk. I didn't use the cardboard reinforcement with this frame.

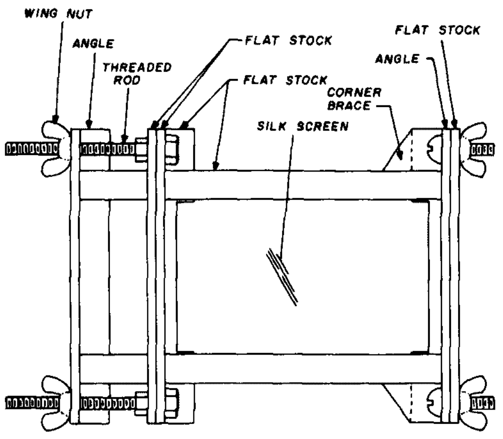


fig. 2. The metal screen printing frame. The cardboard reinforcement is used here.

The second kind of frame can be made from ordinary aluminum stock available in most hardware stores. I use 3/4-inch angle and 3/4-inch flat stock. One clamping surface for holding the screen is fixed; the other is moveable. There are two fixed elements, the second being used for application of the required tension. One of the fixed angle pieces is reinforced to the flat bars with corner braces for additional stiffness. The moveable angle is coupled to the second angle piece with 1/4 x 20 inch threaded rod; wing nuts are used for adjusting tension on the silk.

The one fixed angle element and the moveable one are arranged so that the two ends of the screen, supported by cardboard as explained in the next section, can be clamped tightly to the two members. In this way, ample tension can be applied to the screen for use in printing (see **fig. 2**).

screen preparation

The silk screen is prepared by washing, again with the cleanser, and thorough rinsing. A monofilament nylon screen material of the finest possible mesh is best and will give the finest resolution and minimum problem from etch-through resulting from blockage of ink penetration by the screen material itself. The screen must be stretched as tightly as possible when used, since only then can sufficient contact of the pattern and the copper be achieved, minimizing "rununder."

To protect twisting the thread pattern of the screen material, use cardboard bracing strips on each tension edge, leaving enough silk to wrap around the strip. The silk can then be stapled to the cardboard strip and the combination tacked on the frame or clamped as required. This way the stress can be distributed uniformly on the silk.

using printing film

Since there are two possible ways the screen master can be used, each method is considered separately. I have found orange printing film to be useful and easy to prepare for simple circuits. In using it, one simply *marks off and removes* narrow ribbons of film to form conductors, lifting them from the backing material. The material removed represents a current path. Care should be taken to minimize the cutting of the backing, a plastic, nylon-type material, as the transfer of the film to the silk is most easily accomplished if the film has been cut through without scoring the backing.

I have made some simple tools for preparing the film. One type, for cutting conductor paths, consists of two halves of a double-edged razor blade mounted on opposite sides of a piece of used copper-clad (see **fig. 3**). This will cut both sides of a conductor path at one time, and help in making sure that the length of the cut is correct. These cuts can extend to about a hundredth of an inch into an adjacent pad or across an intersecting path to simplify the removal of the material. This ribbon is then picked up with an Xacto™ knife or a pin and removed. Pads can be cut with a tool made by taking a short length of 1/4-inch rod, center-drilling it on a lathe or drill press, and cutting down its outside diameter to the size of the pad required.

When the pattern has been prepared, it may be attached to the silk screen material by stretching the screen tightly over the pattern and patting the screen with a piece of cheesecloth wetted with lacquer thinner (use a gentle push, not a sliding motion). You will be able to see where the attachment is satisfactory.

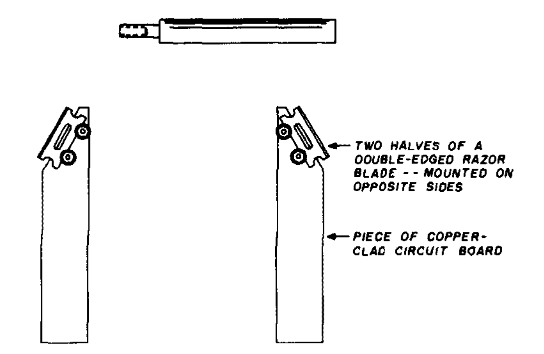


fig. 3. Special tools for use with printing film.

You will want to go back *and redo* any imperfectly imbedded areas. When the combination has dried completely, carefully peel off the backing, resticking if required.

photo-sensitized silk screen

The silk can also have the required pattern applied to it by the use of a photographic sensitizing technique. The sensitizer I have used is the Hunt Manufacturing Company Printing Photo Emulsion Kit No. 4533. This contains two components which are mixed just prior to use. Instructions are provided with the package. A leaflet on screen printing is also available.

To prepare the photo screen, mount the screen material on the frame you have chosen and apply the mixed sensitizer in a thin, smooth layer on both sides of the screen. (You can expedite drying by blowing *cool* air from a hair dryer on to the screen.) After mixing, handle the coated screen in semi-darkness only. If the image to be transferred is closest to the back of your image master, you expose with that surface adjacent to the sensitive surface and expose through it. (Lift-it masters are exposed from the top, whereas drafted masters will be exposed from the bottom; see **figs. 4** and **5**). The master should be between the light and the sensitive layer, and the image as viewed from the top should be as *required. A transparent cover* should be placed on top of the master and weighted so as to assure close contact between the master and the screen. (I use a No. 2 photoflood in a reflector about 14 inches from the work, for about six minutes.) The exposed silk is then washed and rinsed immediately.

inking

After the circuit board has been scoured and prepared for use, and the screen with the appropriate pattern is in tight contact with it, the inking can be begun. The ink must be reasonably thick, yet it must spread through the open areas of the screen. At the same time it must be able to be completely removed from the stencil screen without leaving residues or damaging the screen. It must dry "hard" — that is, it must, after drying, be resistant to the etchant.

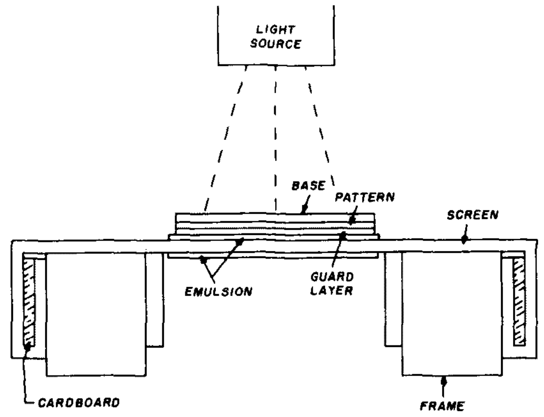


fig. 4. Arrangement for preparing a screen master from a Lift-it of a photo copy of a circuit.

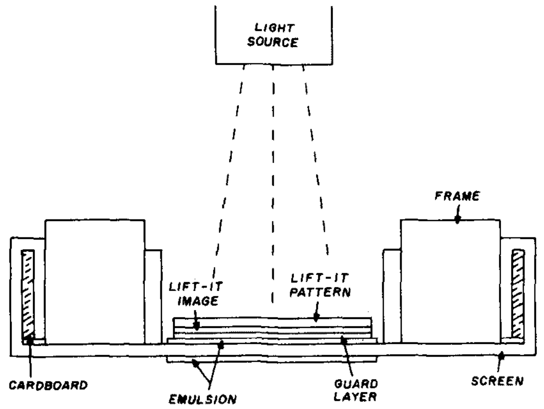


fig. 5. Arrangement for preparing a screen master from a board drawing positive master.

A bead of ink is spread along the short length of the circuit to be printed, and then spread along the image of the circuit. I use a piece of Plexiglass™ or other transparent acrylic as a spreading tool. It should be an inch wide or wider, and can be wide enough to cover the entire width of small boards. (All burrs and sharp edges should be removed from the spreader prior to use. To re-use, simply peel off the dried ink.) Acrylic inks such as the Hunt Permanent Acrylic Screen Printing Ink or the Liquitex Permanent Acrylic Ink are suitable.

After the printed board is dry, the image can be touched up by usng pin or a needle to repair breaks,

or an Xacto knife to scrape away any run-unders that may have occurred. I usually use a hair dryer *with heat* to speed the drying in this phase of production.

I print as many boards as I need in rapid succession and then wash out the screen master with a thorough spray of water. (Printing inks are soluble in water until they dry; after drying, they become impervious to water but can be peeled or scraped off.)

initial artwork preparation

When the photographic screen method is used, it is necessary to work from some kind of master. These masters may be those printed in a magazine (either positive or negative) or some you have prepared from any of the various commercial materials. Each approach is discussed here.

A complex circuit or one available as a circuit pattern in an article can be made into a screen by combining a photocopy of the layout with the silk-screen process. The photocopy can provide increased contrast, if necessary, and eliminate the need to cut the magazine. If the original is positive, make a Lift-it from the photocopy and use it to expose the screen. If the original is a negative, make the Lift-it copy and then print it directly onto a piece of high-contrast 4 x 5 inch cut film. This will give a positive that can be used to sensitize the screen again. (With a negative, the print may be made directly from the Lift-it to the board using the Hunt preparation if you prefer. This works, particularly if the hardener is used as described later.)

I make some of my masters on tracing vellum using extremely thin transfer materials such as those made by Vector. Ruled India ink lines are suitable for conductors. Transfers used for pads and IC sockets finish the circuit layout. The result is a simple, direct step-by-step process.

washing and etching

Washing must be done at several points in this process. The boards should be washed thoroughly and carefully after scouring.The screen material should be washed thoroughly from both sides to remove any sizing and acrylic ink. With the photo-sensitization process, it is spray-washed from both sides to remove the filler from the pattern.

With both the exposed photo-screen and the inked screen, I use a discarded spray bottle for washing, which must be done immediately after completion. A fine but fairly hard spray is best.

I generally use ferric chloride as an etchant. Either plastic or glass trays may be used with it; I use Pryex™ glass trays so I can heat the etchant and thereby speed the operation. My heater is an electric plate warmer with two switches added to the line cord, one with a diode connected across its points for the convenience of two heat levels.

After etching, the board should be washed thoroughly. You'll find the ink softened enough to peel off, leaving the copper with the dull appearance of cuprous oxide. If you wish to apply tinning solution, the copper must be made bright once again by the use of 600 grit paper used lightly as needed and scouring. Hardened ink can be dissolved in lacquer thinner.

hardening

The photo-emulsion image on the screen can be hardened by treating it with Hunt's "Permanizer"™ No. 4529. The developed and dried image on the silk screen is painted with this material, and the combination dried with cool air. The use of a water spray wash with cold water once again opens the mesh where the pattern is.

acknowledgement

I am deeply indebted to A. L. Spizzo of Hunt Manufacturing Company for his assistance in solving various technical problems I have encountered.

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ham radio

DX FORECASTER

Garth Stonehocker, KØRYW

winter DX

The winter DX season is here. One characteristic of winter is a steep rise in the daily MUF peak followed by an early decline to a deeper predawn minimum. This makes for shorter daytime DX operating time in the higher HF bands, but for more nighttime DX on the lower frequency bands. Signal strengths are higher because of lower absorption of energy and less propagated or local atmospheric noise (by this time of year, thunderstorms are fewer and more distant).

Absorption is a result of the loss of energy from the signal as it collides with ions on its path through the D region (about 100-120 miles, or 60-80 km, above the earth). How much energy is absorbed per transit of the D region depends on the location of the sun, and is a function of cosine X, the zenith angle to the sun. Maximum absorption occurs at the subsolar point (directly under the sun); absorption decreases as the signal transit moves away from the subsolar point in any direction. In our winter the subsolar point moves down to 23 degrees south latitude, resulting in less absorption. At the same time the earth is closer to the sun by 2 percent. The net result is still less absorption in winter. The degree of absorption is related to and follows the changes in the ultraviolet output of the sun. (It takes slightly over 8 minutes for a change on the sun to begin affecting our ionosphere.) A measure of this is the daily solar flux at 2800 MHz recorded in Ottawa, Canada, and broadcast at 18 minutes after the hour by radio station WWV. Another source of absorption, caused by increased particle influx during geomagnetic storms, occurs on propagation paths through or along the auroral zone (60-80 degrees latitude). An indication of this cause is an increase in the geomagnetic K (greater than 4) and A (greater than 30) indices, also broadcast from WWV.

On any propagation path, absorption increases with the number of transits of the D region and also varies inversely with frequency. Therefore in working DX it pays to use the higher frequency bands to obtain more distance per hop (resulting in fewer transits) and less signal loss. This is why we generally think of 6, 10, or 15 meters for DXing. But in winter, we have the opportunity to work DX on the lower frequency bands with less QRN and lower signal loss than at any other time of the year.

Lower signal loss is something to look forward to, but you can't count on it. Sometimes in winter, signals are poor for several days at a time. This is caused by anomalous absorption, which will be discussed in next month's column.

last-minute forecast

The low HF bands, 160 through 30 meters, are expected to be the favored bands of operation during the first two weeks of November, with higher bands providing the best DX during the last two weeks of the month. The solar radio flux should be about the same as last year's values, yet higher than it's been in the last month or two. Some possibility of recurrent geomagnetic storms still exists, with greatest probability of occurrence on November 4, 9, 14, 18, and 28. Remember: even though disturbances affect signal strength and produce fading conditions for some paths, conditions on other paths may actually improve.

November is the month during which numerous meteor showers occur. Shower activity should begin on October 26 and last until November 22. A shower maximum of ten per hour is expected during the Taurids meteor shower from the 3rd through the 10th. Lunar perigee is on the 20th; full moon is the 8th.

A total eclipse of the sun will occur on November 22 and 23 in the south Pacific, starting at 2013 UT in the Philippines and New Zealand, traveling east to Antarctica, and ending at 0133 UT. You might want to schedule some contacts with ZL and KC4 land for some unusual DX.

band-by-band summary

Ten, fifteen, and twenty meters will be open from morning to early evening almost every day, and to most areas of the world. The openings on the higher of these bands will be shorter and will occur closer to local noon. Transequatorial propagation on these bands will more likely occur toward evening during conditions of high solar flux and a disturbed geomagnetic field. Absorption effects are not too noticeable.

Thirty and forty meters will be useful almost 24 hours a day. Daytime conditions will resemble those on 20 meters. Skip distances and signal strength may decrease during midday on those days that coincide with high solar flux values. Nighttime DX will be good except after days of very high MUF conditions and the winter anomaly. The usable distance is expected to be somewhat greater than that achieved on 80 at night.

Eighty and one-sixty meters are the nighttime DXer's bands. The bands open just before sunset and last until the sun comes up on the path of interest. Except for daytime short-skip signal strengths, high solar flux values don't affect these bands much. The anomaly will affect day and night signal strength on some days.

ham radio

NOVEMBER

WESTERN USA

GMT	PST	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)
0000	4:00	20	30	15	10	15	10	10	20
0100	5:00	20	30	15	10	15	10	10	20
0200	6:00	20	30	20	10	15	15	10	20
0300	7:00	20	30	20	10	15	15	10	20
0400	8:00	20	30	20	10	15	15	15	20
0500	9:00	20	*30*	20	15	15	20*	15	20
0600	10:00	*20*	*40**	*20*	15	*20*	20	15	30
0700	11:00	*30*	*40*	*20*	*15*	*20*	*20*	*20*	30
0800	12:00	*30*	*40*	*20*	*15*	*20*	*20*	*20*	*30*
0900	1:00	*30*	*40*	*20*	*20*	*20*	*20*	*20*	*30*
1000	2:00	*30*	*40*	*20*	*20*	*20*	*20*	*20*	*30*
1100	3:00	*30*	*40*	*30**	*20*	*20*	*20*	*20*	*30*
1200	4:00	*30*	*40*	*20*	*20*	*20*	*20*	*20*	*30*
1300	5:00	*30*	30	*20*	20	*20*	*20*	*20*	*40*
1400	6:00	*40*	30	20	20	*20*	*20*	*20*	*40*
1500	7:00	*40*	30	15	20	*20*	*20*	*20*	*40*
1600	8:00	*40*	20	15	20	*20*	*15*	*20*	*40*
1700	9:00	*40*	20	15	20	15	15	*20*	*30*
1800	10:00	*40*	20	15	20	15	15	15	*30*
1900	11:00	*40*	20	15	15	15	15	15	30
2000	12:00	*40*	20	15	15	15	15	15	20
2100	1:00	30	20	15	15	15	15	15	20
2200	2:00	30	20	15	15	15	10	15	20
2300	3:00	30	20	15	15	15	10	15	20

MID USA

GMT	MST	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)	CST
0000	5:00	30	30	15	15	15	10	15	20	6:00
0100	6:00	30	30	15	20	15	15	15	20	7:00
0200	7:00	*40*	30	20	20	15	15	15	20	8:00
0300	8:00	*40*	30	20	20	15	15	15	20	9:00
0400	9:00	*40*	30	20	20	15	15	15	20	10:00
0500	10:00	*40*	*30*	20	20	*20*	20	20	30	11:00
0600	11:00	*40*	*40*	*20*	20	*20*	*20*	20	30	12:00
0700	12:00	*40*	*40*	*20*	*20*	*20*	*20*	*20*	*30*	1:00
0800	1:00	*40*	*40*	*20*	*20*	*20*	*20*	*20*	*30*	2:00
0900	2:00	*30*	*40*	*20*	*20*	*20*	*20*	*20*	*30*	3:00
1000	3:00	*30*	*40*	*20*	*20*	*20*	*20*	*20*	*30*	4:00
1100	4:00	*30*	*40*	*20*	*20*	*20*	*20*	*20*	*30*	5:00
1200	5:00	*20*	*40*	*20*	*20*	*20*	*20*	*20*	*40*	6:00
1300	6:00	20	30	*20*	15	*20*	*20*	*20*	*40*	7:00
1400	7:00	20	30	15	15	*20*	*20*	*20*	*40*	8:00
1500	8:00	20	30	15	15	*20*	*20*	*20*	*40*	9:00
1600	9:00	20	30	15	15	*20*	*20*	*20*	*40*	10:00
1700	10:00	20	20	15	10	15	15	*20*	*30*	11:00
1800	11:00	*30*	20	15	10	15	15	15	*30*	12:00
1900	12:00	*30*	20	10	10	15	15	15	*30*	1:00
2000	1:00	30	20	10	10	15	15	15	20	2:00
2100	2:00	30	20	10	10	15	15	15	20	3:00
2200	3:00	30	20	15*	15	15	10	15	20	4:00
2300	4:00	30	20	15	15	15	10	15	20	5:00

EASTERN USA

GMT	EST	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (Caribbean S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)
0000	7:00	30	30	15	15	15	15	15*	20
0100	8:00	40	30	15	20	15	15	15	20
0200	9:00	*40*	30	20	20	15	15	15	20
0300	10:00	*40*	30	20	20	15	15	15	20
0400	11:00	*40*	30	20	20	20	20	15	30
0500	12:00	*40*	*40**	20	20	*20*	20	20	30
0600	1:00	*40*	*40*	*20*	*20*	*20*	*20*	20	*30*
0700	2:00	*40*	*40*	*20*	*20*	*20*	*20*	*20*	*30*
0800	3:00	*40*	*40*	*20*	*20*	*20*	*20*	*20*	*30*
0900	4:00	*30*	*40*	*20*	*20*	*20*	*20*	*20*	*30*
1000	5:00	*30*	*40*	*20*	*20*	*20*	*20*	*20*	*30*
1100	6:00	*30*	*40*	*20*	*20*	*20*	*20*	*20*	*40*
1200	7:00	*20*	*30*	*20*	15	*20*	*20*	*20*	*40*
1300	8:00	20	30	*15*	15	*20*	*20*	*30*	*40*
1400	9:00	20	30	15	15	*20*	*20*	*20*	*40*
1500	10:00	20	30	15	15	*20*	*20*	*20*	*40*
1600	11:00	20	20	15	15	*20*	*20*	*20*	*40*
1700	12:00	20	20	15	10	15	15	15	*30*
1800	1:00	20	20	15	10	15	15	15	*30*
1900	2:00	30	20	15*	10	15	15	15	*30*
2000	3:00	30	20	10	10	15	15	15	20
2100	4:00	30	20	10	10	15	15*	15	20
2200	5:00	30	20	15*	15	15	10	15	20
2300	6:00	30	20	15	15	15	10	15*	20

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during "normal" hours.

*Look at next higher band for possible openings.

Joe's done it again — a column of interest to HF'ers and VHF'ers alike. — Ed.

VHF/UHF WORLD

Joe Reisert W1JR

high dynamic range receivers

Mention high dynamic range and you'll really get a discussion going. Everyone has a story and a solution. We'd all like to believe that our receivers or transmitters are always clean, and that any splatter or other obnoxious noise has to be coming from somebody else's poor receiver or dirty or overdriven linear amplifier. Because all aspects of the situation, both on the transmit and receive side, are seldom separated, the problem is rarely resolved.

A few years ago, after lots of arm-twisting by Jim Stitt, WA8ONQ, I tackled this dilemma. The main goal was to improve Jim's 6-meter receiver sufficiently so he could be sure it wasn't the culprit in these situations. Then his station could be more competitive in the VHF contests and he'd be able to operate alongside the strong local transmitters — assuming *they* were also clean (more on this later).

The end product was a high dynamic range 6-meter receive converter with an extra transmitter LO (local oscillator) output. Since this month's *ham radio* emphasizes receivers, I decided to discuss this subject in some detail and examine some of the problems, limitations, and solutions for such a design. Typical recommended circuits for a 6-meter receive converter will then be shown.

high dynamic range

What is high dynamic range? One answer is that it describes transmitter or receiver design that allows copying weak DX signals in close proximity to strong signals. It sounds simple enough, but how strong is that local? Let's assume that a big 6-meter station is only 1 mile (1.6 km) distant and runs the legal limit of 1.5 kW PEP output through a feedline with a 1 dB loss to a 10 dB gain antenna pointed at your receiving antenna. If you're using a dipole receiving antenna broadside to this signal and have no feedline loss, the signal at the input to your receiver will be approximately 0 dBm or 1 milliwatt, about 130 dB above the noise floor in a typical VHF receiver! If you also have a 10 dB gain antenna, the signal received (when aimed at this source) will be +10 dBm (10 milliwatts) — more power than is used for the LO in most Amateur receivers! If the distance between stations is doubled, the signal will drop by 6 dB but still be quite respectable.

Recently I tested the dynamic range of a well-designed Amateur 6-meter converter that uses a single JFET preamplifier and a standard level (+7 dBm or 5 milliwatts) DBM (doubly-balanced mixer). When two equal signals of −20 dBm (23 millivolts or 10 microwatts) were present at the input to the converter, spurious signals or IMD (intermodulation distortion) were generated and only 30 dB below the desired outputs. This is hardly high dynamic range! These spurs or IMD products usually appear as sidebands or additional signals spaced equally above and below the normal signals by the difference between the input frequencies (see **fig. 1**). When the IMD gets worse, additional spurs appear alongside the first sidebands as is also seen in the photo in **fig. 1**.

All is not lost. The antennas can be part of the solution. If you use a directional antenna with a clean pattern, moving your antenna back and forth can place a null on a strong signal.[1] If overload is still present, one solution is to place an attenuator at your receiver input (more on this later); on VHF, especially on 6 meters, where the local ambient noise is usually high, the weak signal will still be good copy while the local (interfering signal) will have been "knocked" down.

is the receiver at fault?

Before proceeding, it may be worth mentioning the transmitter. Typically speaking, it would be desirable for all Amateurs to transmit a clean signal. But what is a clean transmitted signal? Typical Amateur linears call for IMD products to be at least 30 dB below the desired signals. However, if the received signal is 130 dB above the noise, 30 dB IMD isn't going to be much help 3 kHz from the other station's frequency. You'll just have to QSY further away.

Therefore, before you accuse the other station of "hitting it too hard," perform one simple test. *First observe the splatter several kHz away* on a relative basis or on your receiver "S" meter. Next, place an appropriate attenuator (10 dB recommended, see **fig. 2**) at the input to your receiver or converter (or use the internal attenuator if one is an integral part of your receiver) and then recheck the splatter level. If the received level drops by approximately the insertion loss of the attenuator, the transmitter is the culprit and the transmitting station is

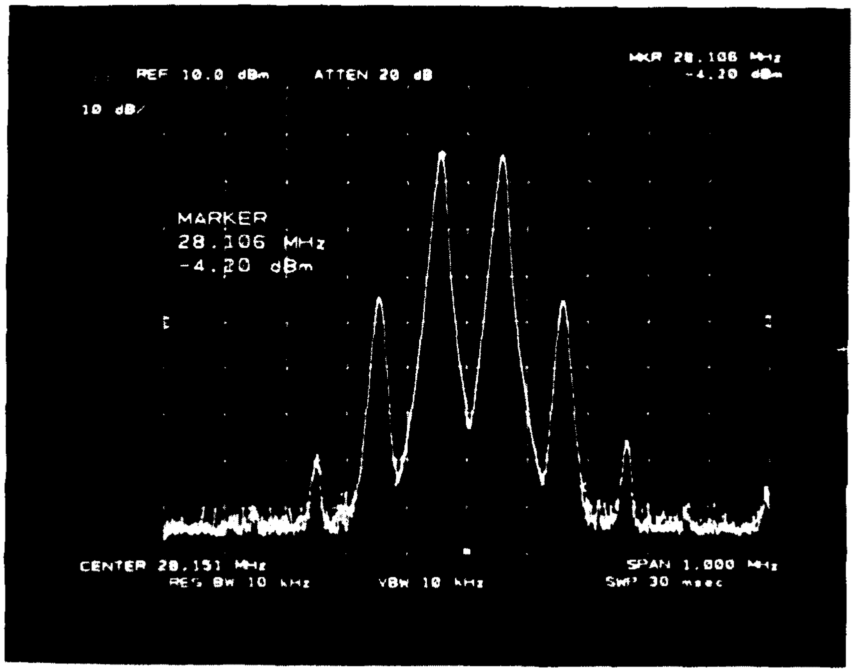


fig. 1. A spectrum analyzer shows the 28 MHz IF output of a typical 6-meter converter as described in the text. Input signals are approximately 50.1 and 50.2 MHz at − 20 dBm each.

either overdriving its equipment or your frequency is just too close for comfort. However, if the splatter drops by more than the attenuator value (it could be up to 3 times less!), your receiver is surely part of the problem.

Assuming that the problem is the receiver (maybe in some future column we'll examine transmitter and power amplifier requirements in greater detail), there are design approaches that will enhance receiver performance.

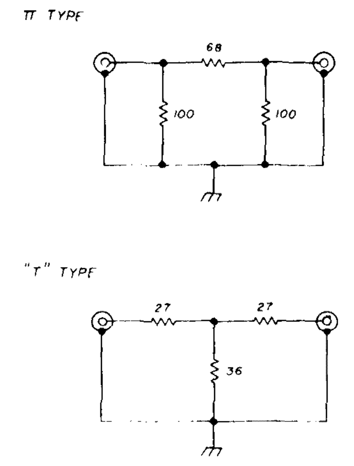


fig. 2. Typical 10 dB attenuator pads for testing or improving the dynamic range of a receiver.

general receiver design requirements

The old saw "If you can't hear them, you can't work them" still applies to high dynamic range receivers. Low noise figure, sufficient RF selectivity to reject images and undesired out of band signals plus a clean local oscillator are still required.

In order to obtain low noise figure, a preamplifier is usually required ahead of the first mixer. Herein lies the problem. Any gain ahead of the mixer will decrease *the dynamic range*. Therefore, the preamplifier gain must be kept as low as possible, consistent with obtaining the desired noise figure. The desired or total system noise figure can be determined from:

$$NF\ (total) = NF1 + (NF2 - 1)/G1 + (NF3 - 1)/G1G2,\ etc. \quad \textbf{(1)}$$

where G is gain and NF is noise factor (a numeric-not dB) for each stage in succession.

For example, a 3 dB noise figure preamplifier with 9-10 dB of gain feeding a mixer with a 10 dB noise figure will yield an overall system noise figure of less than 5 dB. Fortunately, we can live with lower sensitivity receivers, especially on 6 meters, where a noise figure less than 5 to 10 dB is usually wasted since the typical ambient noise is usually very high.* At 2 meters the problem is more acute, but the level and number of signals are usually somewhat less of a problem. For the computer-minded Amateur, a computer program is *available for* **eq. 1** so that you can quickly iterate various combinations of gain and noise figure to determine your own optimum case.[2] Before leaving the subject, remember that a preamplifier must have high output power capability in order not to distort the signals prior to the first mixer (more on this later).

For good performance you need adequate RF selectivity ahead of the mixer, which means additional losses that further increase the noise figure. Again, these losses can be handled (as we shall soon see) by the proper choice of RF filter characteristics and by optimizing the location of the filters in the receiver chain.

Let's not forget the choice of IF and its effect on selectivity, images and spurious responses. For 6 meters I personally favor a 28 MHz IF with a 22 MHz LO, rather than a 14 MHz IF with the 36 MHz LO used on some of the older converters. Spurious signal analysis reveals that a 28 MHz IF is slightly less susceptible to "birdies."[3] Also, a 28 MHz IF is far less likely to respond to IF breakthrough. The latter term refers to leakage at the IF frequency that permits normal signals in this spectrum to also be received. The 20-meter band is a good example since propagation yields signals of high intensity, especially during the days when 6 meters is hot. Although 10-meter IF breakthrough can be a

*Everything's relative — Joe, who also operates 160 meters, would probably agree that in contrast, 6 meter ambient noise is low — **Editor**

problem, the number and strength of stations present is usually less, especially below 28.3 MHz.

High dynamic range mixers that require moderate LO power (10-100 milliwatts) are usually required. Also the LO must be very clean with low phase noise (more on this later) and should be followed by an amplifier to boost the level high enough to adequately drive the mixer.

Finally, if the overall system noise figure is to be realized, the mixer must usually be followed by a low noise figure postamplifier with a high dynamic range. The IF receiver should also have high dynamic range and a moderate (10 dB or so) noise figure.

preamplifiers

Surely the preamplifier is one of the most important aspects of a good receiver. However, obtaining high dynamic range and low noise figure simultaneously and with a reasonable input and output VSWR is difficult. Devices (transistor, FET, etc.) with low inherent noise figure are common. However, increasing preamplifier dynamic range usually requires increased device current or a device with greater current-carrying capacity. This, in turn, usually increases the noise figure and the overall gain, the exact opposite of the desired effect!

Before discussing different preamplifiers in detail, it may be well to mention the subject of the linearity in an active device. Just because an amplifier is operated in class "A" doesn't mean it is free from distortion. *Every amplifier, regardless of its type and power, has a point beyond which the output signal will no longer be an exact replica of the input signal.* Hence distortion will occur.

Over the years various methods have been devised to measure distortion. The most frequently used test is for 1 dB compression. This is defined as the CW power level where the output signal increases 9 dB for an input power increase of 10 dB. Most class "A" amplifiers can only increase output power by 2-6 dB beyond this level, as shown in **fig. 3**. Amplifiers often are heavily distorting a signal 5-10 dB before it reaches compression levels; consequently, this is not a good point for referencing distortion. Furthermore, some devices are more nonlinear than others, especially when approaching the compression point.

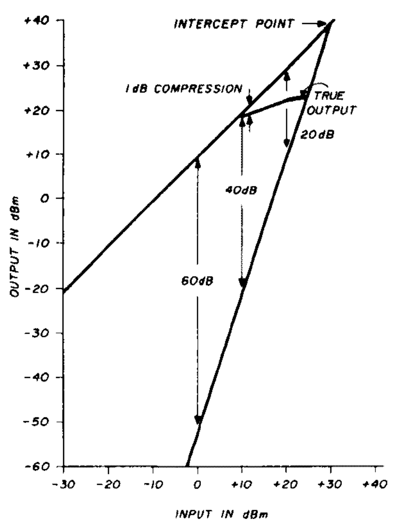


fig. 3. A high dynamic range preamplifier (fig. 6) input versus output level response with typical IMD levels and compression levels.

In 1967, McVay wrote his classic reference paper on the third-order intercept point, a new method of measuring dynamic range.[4] Basically, what this method does is to determine distortion based on a two-signal IMD test performed in a similar manner to that used to specify single sideband linear amplifiers. The third-order intercept point is then determined either mathematically or by use of nomograph (see **fig. 3**). The distortion can then be calculated or read off the nomograph for any power level on any device if the third-order intercept point is known. Suitable nomographs are available in reference 4 and from most commercial amplifier manufacturers.

In most of the work I have done on high dynamic range, I have used the intercept point test method. Several things are immediately apparent. The IMD products increase at three times the rate of the desired output signal level change. Hence, the ratio between the output signal level and distortion will change on a 2 for 1 basis. For example, if the IMD products from two equal level signals are 60 dB below the desired output signal level and the signal level is increased by 1 dB, the IMD products will now be only 58 dB below the desired output level. A 10 dB input signal increase will decrease the IMD difference by 20 dB. This can be seen graphically in **fig. 3**. Therefore, once IMD becomes apparent, it will usually degrade very rapidly, perhaps even on a greater than 2-to-1 basis, with increased signal level! This is common on many active devices whenever the IMD is less than 60 dB below the output levels.

A search was launched for the ideal preamplifier. First a low-gain (12-13 dB) grounded-gate J310 JFET preamplifier was designed (after all, FETs are supposed to have such great dynamic range and low noise figures). The results were fair. On a typical circuit the output compression point was +14 dBm (250 milliwatts). The IMD was down 60 dB for −3.5 dBm (0.45 milliwatts) output per signal for an output third-order intercept point of +26.5 dBm (450 milliwatts). However, the noise figure was over 4 dB with no input matching. When input noise figure matching was added, the gain increased and the input impedance match degraded — both detrimental to the desired results. Also, the overall selectivity for this preamplifier was inadequate for the final converter.

Before proceeding with the next preamplifier design, some re-examination was in order. Previous experience with modular circuits led to the conclusion that in a high dynamic range receiver all circuits should have good input and output VSWR at a common impedance such as 50 ohms.[5] This would allow easy interchange between filters, amplifiers, mixers, and LOs, as well as facilitate any future improvements or changes, especially when new or improved devices became available.

With this in mind, the search for a low-noise high dynamic range pre-

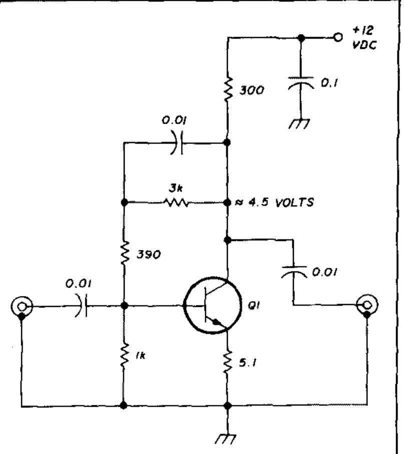


fig. 4. A high dynamic range 14 dB gain preamplifier using RF feedback for wide bandwidth. (See text for complete specifications.)

for −60 dB IMD were −4 dBm (0.4 milliwatts) for an output intercept point of +26 dBm (400 milliwatts). Higher output power could be obtained with still higher I_C. Unfortunately, when high feedback and high I_c are used, the noise figure also increases. In this case the noise figure was already about 4 dB for an I_c of 25 mA. Adding more current or a 4:1 output transformer would have resulted in an undesirable increased noise figure and equally undesirable increased gain.

Not being totally content with this amplifier, I tried one of the less expensive (approximately $8.25 each) broadband hybrid amplifiers, a Motorola MWA 130, which exhibits a +19 dBm (95 milliwatts) compression point. For 60 dB IMD, the outputs were +4.77

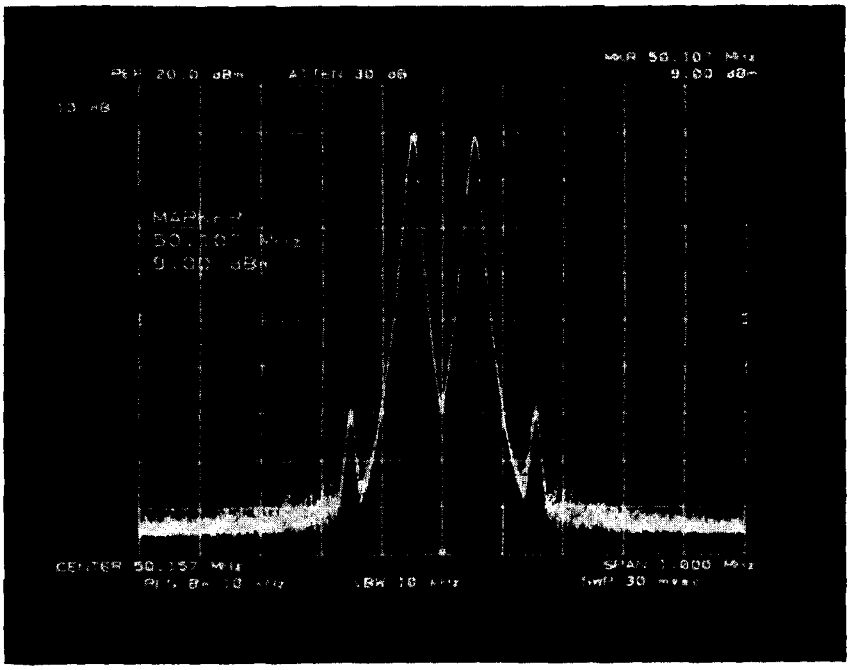


fig. 5. A spectrum analyzer photograph of the output of the high dynamic range amplifier (fig. 6.) with two equal level input signals at approximately 50.1 and 50.2 MHz at 0 dBm each.

amplifier with good input and output VSWR began. A preamplifier was designed around the 2N5109 transistor, a CATV favorite, using shunt and series feedback to obtain a matched input and output impedance (**fig. 4**). The VSWR was less than 1.5:1 from 1 to 70 MHz range while the IMD was

acceptable but only with high (25 mA) I_c (collector current). Typical outputs dBm (3 milliwatts) for an output intercept point of +34.8 dBm (3 watts), a substantial improvement over the home-brew circuit. However, the current drain was 60 mA and the noise figure was about 6.5 dB at 50 MHz,

similar to the performance of the 2N5109 circuit just described, when its current was raised to the same level. Also, the gain — over 15 dB — was too high for this application.

I finally tested one of my favorite preamplifiers, a single transformer lossless feedback type using a common base circuit similar to the one designed by Norton.[6] Although it is more complex to construct, the results are well worth the effort. Using a medium gain (9 dB) configuration, the output power and IMD were outstanding, provided the emitter current was moderate (17 mA). Output compression was typically +20 dBm (100 milliwatts). IMD was down 60 dB for +9 dBm (8 milliwatts) output, for an output intercept point of +39 dBm (8 watts)! The typical IMD versus input and output for this circuit is shown in **fig. 3** and a typical two-tone spectrum display is shown in **fig. 5.** As a bonus, if the preamplifier is properly constructed, the bandwidth is greater than 1.8 to 200 MHz with a 2:1 maximum VSWR and 10-150 MHz for a 1.2:1 VSWR! Truly this was the circuit I was searching for (**fig. 6**).

A big key to the success of a high dynamic range preamplifier is the type of transistor chosen. Many RF devices will work well but not always have the same noise figure, bandwidth, or IMD. In the lossless feedback case, the noise figure was typically 1.5 to 2 dB maximum when using the NEC NE41632B transistor, but a 2N5109 had a noise figure of 2.5 to 3 dB in the same circuit. In addition, previous work showed that the most linear transistors were those which were specifically designed for CATV and class "A" linear operation with a *very constant DC current gain (h_{fe}) over a wide range of collector current*. In the CATV business, which is particularly interested in IMD, these devices are frequently referred to as large area multiple emitter structures. The NE41632B and the 2N5109 are both included in this category. (For those who do not have easy access to the NE41632B transistor or the balun core shown in **fig. 6B**, I have made arrangements for PROTO-FAB,

74 Wedgemere Drive, Lowell, Massachusetts 01852, to provide them at a nominal cost. Write them for price and delivery information.)

Caution: This circuit has been modified and has a higher dynamic range than the original Norton circuit. However, his original circuit is patented (U.S. Patent No. 3,891,934, issued June 24, 1975, to David E. Norton and Allen F. Podell). Therefore, any attempt to duplicate this circuit for profit may violate the rights held by the Anzac Division of Adams Russell, Inc.

RF filtering

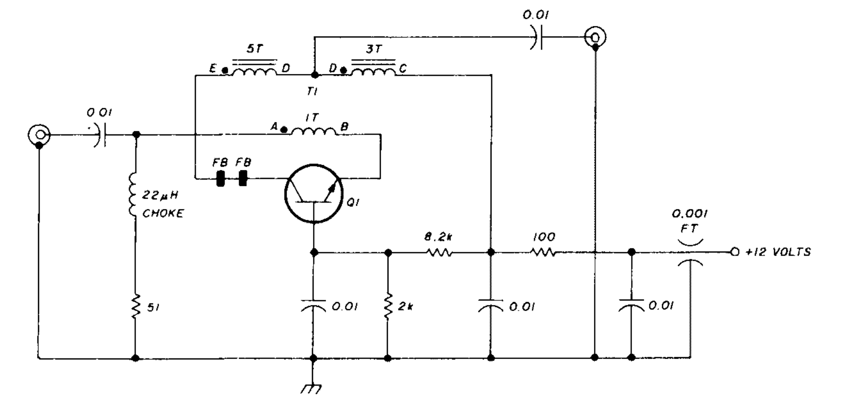
It goes without saying that high dynamic range cannot be obtained if spurious frequencies or high power out-of-band signals are present in the receiver. Hence RF filtering is very important. It was pointed out in a prior article that the type of input filtering chosen can lessen the chances of destruction from HF signals or lightning entering the first preamplifier of a receiver.[5]

In my August column I discussed the problems of multiple pole filtering such as VSWR distortion and increased losses. Hence it was decided to use a simple low-loss single pole bandpass filter with a pseudo-highpass response at the input to the receiver.[7] In this case a 5 MHz bandwidth was chosen because it would allow reception of 48 MHz European video carriers as well as 52 MHz VK/ZL DX with little degradation at either frequency, but still reject other services. This filter has a nominal insertion loss of 0.75 dB, less than a multi-section type. The schematic is shown in **fig. 7**, with its typical frequency response in **fig. 8**. The input filter chosen doesn't have sufficient out-of-band rejection by itself. Hence a post filter (**fig. 9**) with the same bandwidth (5 MHz) but higher insertion loss (2 dB typical) is required. Its frequency response is shown in **fig. 10**. The filter topography may be somewhat new; it was developed by this author and William K. Talley while at the Mitre Corporation in an effort to obtain a symmetrical attenuation versus frequency response.[7] This filter design is available in computer-aided design form in reference 2. Since this filter is placed after the preamplifier, the loss has a minimal impact on overall system noise figure. As a bonus, the extra insertion loss of the filter will improve the overall system dynamic range accordingly.

mixers

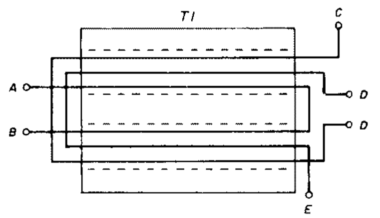
In prior work I had experienced poor dynamic range with the more common Amateur type of mixers such as dual-gate MOSFETs and single-ended bipolar and JFET mixers. However, despite conversion loss, DBMs have always performed very well in my circuit designs.[5,8,9] Hence, when striving for high dynamic range, I decided from my prior experience that DBMs "are the only way to fly." They are easy to drive with a reasonable 50-ohm impedance match at all ports, a goal stated earlier. Also, because of their balanced structure, they tend to cancel any AM present on the local oscillator, a problem which is particularly prevalent if phase-locked LOs are used.

However, when striving for high dynamic range, DBMs must be treated properly. Attenuator pads on the RF and LO ports are a must to terminate undesired mixer generated products and the LO as well as to terminate any



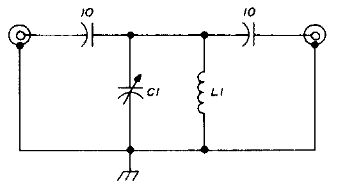
item	description
FB	ferrite bead. Type not critical
Q1	NEC NE41632B or equivalent. (See text for supplier.)
T1	1:3:5 turns ratio transformer wound with No. 32 AWG enameled wire on a Fair-Rite Products type 2843002402 balun core. (See text for supplier.) (Patent warning here.)

fig. 6A. A low-noise wide bandwidth high dynamic range preamplifier using transformer coupled lossless feedback. (See text for further specifications.)



Construction details.
1. First wind one complete turn of No. 32 AWG enameled wire through the balun core from point A to B.
2. Wind three complete turns of the same wire from point C to D. To aid in identification when the transformer is completed, place a small knot in the wire at starting point C.
3. Wind 5 complete turns of the same wire from points D to E.
4. Strip end points D, twist together, and solder.
5. Strip and tin the remaining wires and connect them to the proper point in the circuit. To prevent connecting the transformer incorrectly, leave the small knot on the B+ lead intact until the final wires are soldered in place.

fig. 6B. Balun core is available from PROTO-FAB, 74 Wedgemere Drive, Lowell, Massachusetts 01852, at nominal cost.



item	description
C1	2-20 pF variable capacitor.
L1	approximately 0.28 μH. 9 turns No. 24 AWG enameled wire on Micrometals type T50-10 or equivalent toroid.

fig. 7. A recommended input filter for a 6-meter high dynamic range receive converter. (See text for specifications.)

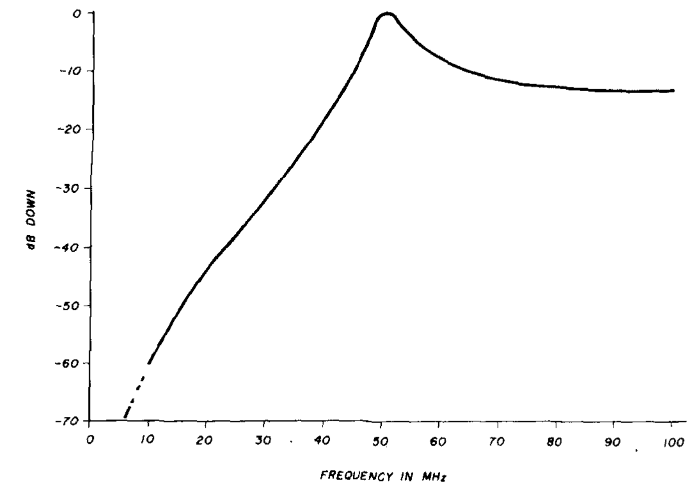
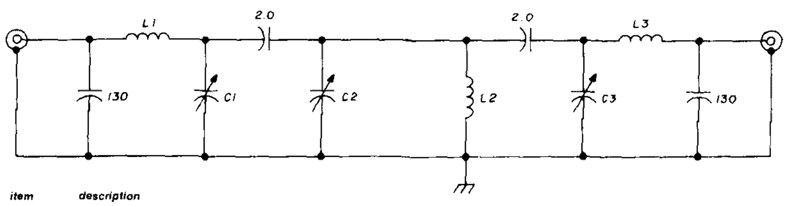


fig. 8. The attenuation-versus-frequency response of the bandpass filter shown in fig. 7.



item	description
C1,2,3	2-20 pF trimmer capacitor.
L1,2,3	approximately 0.42 μH inductor. 11 turns of No. 24 AWG enameled wire on Micrometals type T50-10 or equivalent toroid.

fig. 9. A recommended three-section bandpass filter for a 6-meter receive converter. (See text for specifications.)

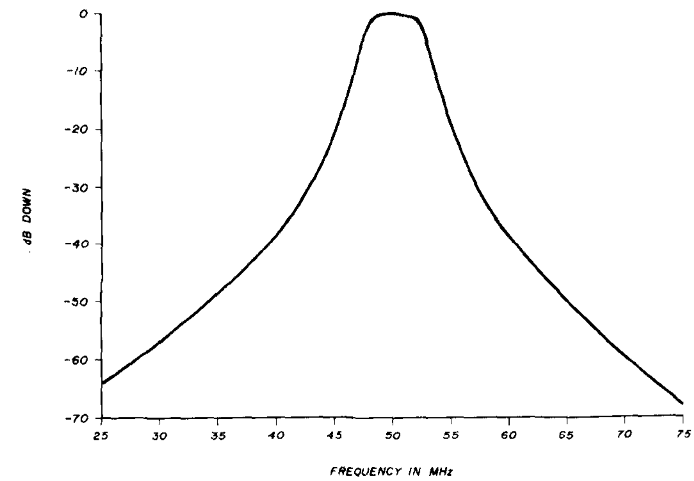


fig. 10. The attenuation-versus frequency response of the 6-meter bandpass filter shown in fig. 9.

in-line filters in their proper impedances.[8] Likewise, a diplexer should be added to the IF output port if low IMD is to be maintained.[8,10]

Many DBMs were tested, with the Mini-Circuits Labs TAK-1H selected as the best overall mixer on a cost-versus-performance basis. For comparison, some of the data taken on this and some other popular DBMs are summed up on **table 1**. The final circuit using the TAK-1H is shown in **fig. 11**. This mixer requires higher LO power (+ 17 dBm or 50 milliwatts) than the more common DBMs usually seen in Amateur equipment, but this is a definite need if high dynamic range is to be obtained. Due to the 3 dB pad on the LO port of the DBM, the LO power required by the overall circuit in **fig. 11** is +20 dBm (100 milliwatts). The RF and LO bandwidth are 2 to 500 MHz. Hence, this circuit has considerably more capability than meets the eye.

A few final remarks on DBMs are in order. Although the so-called high dynamic range mixers (those specified for use with + 17 dBm or 50 milliwatts LO), are recommended, the typical DBMs specified with a +7 dBm (5 milliwatts) LO can be used, but with 5 to 10 dB lower LO and dynamic range. Sometimes *DBMs are not* readily available in single quantity. This can often be handled by getting a group of persons together to buy the minimum quantity. PROTO-FAB, as mentioned earlier, has also agreed to make the TAK-1H DBM available at a reasonable price. Finally, many DBMs are now showing up at flea markets at some pretty good prices, so shop around. You may not find the exact DBM desired but you may be willing to accept slightly lower performance as a compromise. If you adopt the modular approach suggested, it will be easy to upgrade performance at a later date.

local oscillators

My favorite crystal oscillator is the overtone Colpitts.[5,8] The frequency of the LO is determined by the IF chosen, as discussed above. It has great stability and low phase noise, a requirement

table 1. Typical measured data on commonly used DBMs. Input signals are at 50, LO at 22, and IF at 28 MHz.

type	LO (dBm)	input compression (dBm)	output intercept (dBm)	approximate cost	quantity
SBL-1	+7	+2	+14	4.50	(10-49)
MD-108	+7	+4	+16	14.00	(1-5)
SRA-1	+7	+4	+16	11.95	(1-49)
SRA-1H	+17	+12	+22.5	17.95	(5-24)
MHP-106	+17	+13	+22.5	45.00	(1-5)
TAK-1H	+17	+16	+28	19.95	(5-24)
MD-139	+20	+17.5	+29	115.00	(1-5)
RAY-3	+23	+16.5	+24	34.95	(4-9)
SAY-1	+23	+20	+32.5	54.95	(1-9)
VAY-1	+27	+24	+36.5	74.95	(1-9)

of any high-performance receiver. Phase noise, caused by poor design in the phase lock loops employed, is typically poor in many of the transceivers presently available. Phase noise generates noise on the LO, which, in turn, causes strong signals to be heard several kHz away.

As shown in reference 8, the output of this LO is only about +10 dBm (10 milliwatts). Therefore, an amplifier is required if a high level DBM is used. It was decided not to fight the class "C" type of amplifier, but to go class "A" because there would be improved linearity and less possibility of generating LO noise. Since design of the 2N5109 feedback amplifier had already been completed (**fig. 4**), the bias values were modified slightly for use as the LO amplifier. A simple 1/2 λ low-pass filter followed this amplifier to keep any harmonics from reaching the mixer circuitry.[7]

When the preliminary design was completed for WA8ONQ, it was decided to take the LO output through a two-way power splitter for use on both the receive and transmit mixer. This was more power than required by the transmit mixer, and also dropped the output power substantially on the receive side. So a unique connection was made at the output of the oscillator in conjunction with the attenuator usually used at this point.[8] The result is a secondary output sufficient to drive a standard level DBM (+7 dBm) in a transverter similar to those designs in reference 9. If this output is not needed, terminate it with 51 ohms for

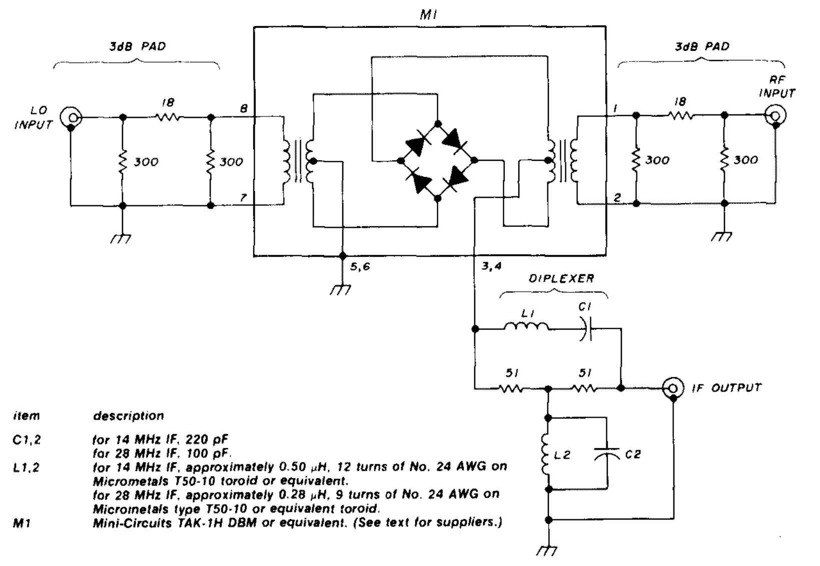
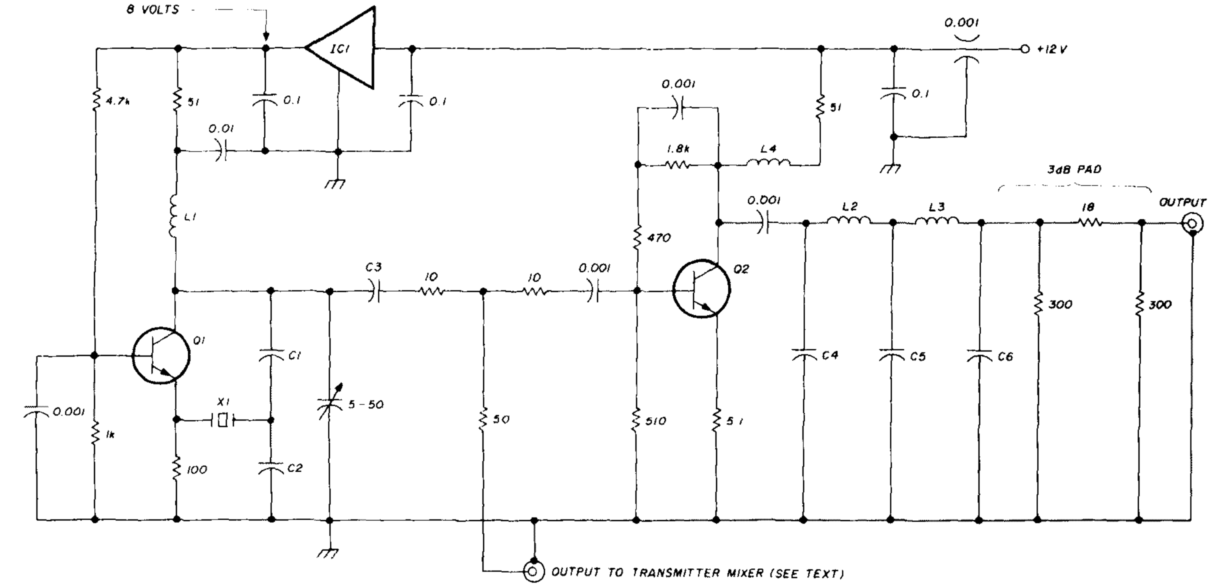


fig. 11. Recommended DBM circuit for a high dynamic range receiving converter with a 14 or 28 MHz IF. (See text for full capabilities.)

possible future use and to insure that the oscillator is seeing the proper match. The final LO schematic is shown in **fig. 12** and delivers +20 dBm (100 milliwatts) at the output connector, the power required by the mixer circuit in **fig. 11**.

postamplification

As already stated, the DBM type of mixer has conversion loss and, therefore, must be followed by a low-noise postamplifier if the system noise figure is to be preserved. The signal levels at this point are about equal to those present at the input to the preamplifier. Therefore, the preamplifier already used is an excellent candidate for this amplifier since it has a good impedance match, low noise figure and high dynamic range. This is also a recommended circuit for HF, 10-meter OSCAR reception, interface with existing VHF/UHF converters, or other applications where moderate gain, low noise figure and high dynamic range are required over a broad bandwidth. In my shack, I have a DPDT coax switch which allows me to bypass the postamplifier when strong signals are present, thereby increasing dynamic range.



item	22 MHz (28 MHz IF)	36 MHz (14 MHz IF)
C1	50 pF	24 pF
C2	200 pF	100 pF
C3	33 pF	20 pF
C4,C6	150 pF	100 pF
C5	300 pF	200 pF
L1	11 turns No. 24 on T50-10 toroid	9 turns No. 24 on T50-10 toroid
L2,L3	10 turns No. 24 on T50-10 toroid	8 turns No. 24 on T50-10 toroid
X1	22 MHz third overtone series mode crystal	36 MHz third overtone series mode crystal
IC1	8 volt, 3 terminal regulator, 76L08 or equivalent	
L4	6.8 μH RF choke	
Q1	2N2222 transistor	
Q2	2N5109 transistor	

(Use a small gear-type heat sink on Q2)

fig. 12. Recommended 22 or 36 MHz LO for a 6-meter high dynamic range receive converter. (See text for output levels.)

IF requirements

At this point it must be obvious that the converter presented has to be followed by a high dynamic range IF receiver if the true capabilities are to be realized. Bypassing the postamplifier or inserting an adjustable attenuator after the receive converter can also help improve dynamic range, but the ultimate limits will probably be limited by the IF receiver. Fortunately, the commercial manufacturers are improving the HF gear that is now being marketed. Furthermore, a converter of the type just described, in conjunction with one of the more modern HF receivers, will definitely outperform any presently available equipment that is solely devoted to the VHF or UHF spectrum.

construction

The 6-meter converter was designed using a modular approach.[5] The final block diagram is shown in **fig. 13**. Each circuit was placed in a cast aluminum box such as the Pomona 2417, Bud CU 123/124 or equivalent with BNC type input/output connectors and feedthrough-type capacitors on the power supply lines. Each box has a piece of double-clad PC board attached to the cover for point-to-point wiring and grounded, as previously discussed in reference 8. The final

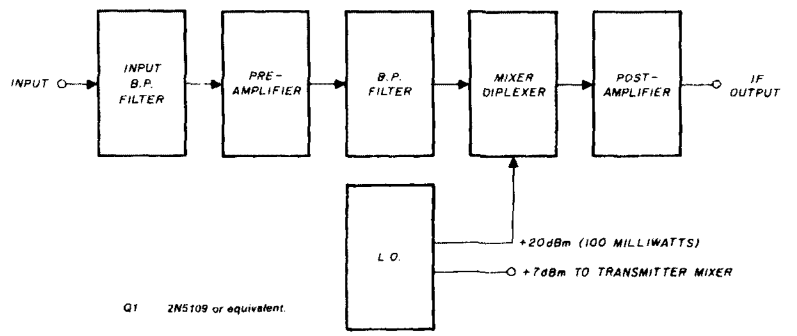


fig. 13. Overall block diagram of a recommended high dynamic range 6-meter converter.

result is a versatile unit with no apparent RF pickup or interaction between modules.

Construction of the circuits is quite straightforward. Leads should be kept short, especially on the filters. High-quality trimmers with low lead inductance such as the ceramic, mylar, or teflon types should be used. The transformer construction hints (**fig. 6B**) for the preamplifier should be followed carefully and the leads on the capacitors, especially on the base bypass, should be kept as short as possible. If a different DBM is used, check the pin designations as some manufacturers use different pin-outs.

tune-up and performance

Very little tuning is required. The input filter can be easily aligned by tuning for maximum signal at 50.1 MHz. The second filter may require more effort. It is best tuned with a sweep setup. However, on the tests I conducted, it was very close to nominal if all capacitors are tuned for maximum signal at 50 MHz in a matched test setup and then inserted in the chain. All the LO requires is to be tuned for maximum output. Properly aligned, the converter described has typically a 5 MHz bandwidth, a gain of about 4 to 6 dB and a noise figure less than 6 dB.

future designs

The state of the art is constantly improving. If a modular approach is used on this receive converter, new or improved circuits can be easily added or changed as prices decrease or parts become available. If you don't change the IF frequency, you'll probably never have to build another LO. If a lower LO power is required, just add an appropriate attenuator on the output. Higher dynamic range mixers are slowly decreasing in price while increasing in performance. **Table 1** can be used as a guide to selection of DBMs.

Finally, our IF receivers must be improved, especially on dynamic range and phase noise. Ultimately, I think that the best receiver will be one that uses a high dynamic range converter directly feeding a narrow bandwidth crystal filter. However, this will require a variable LO and some additional design to prevent phase noise and birdies.

Again, I have rambled on and written a more lengthy column than I intended. However, I feel that the material presented is broad enough in scope and should be worthwhile regardless of frequency. As stated earlier, even using more conventional circuits such as standard level DBMs and JFETs, a substantial improvement can be made over most existing receive converters. After all, the principles discussed are usable at any frequency if time and money are no object!

I hope this material will encourage you and others to try to improve receiver dynamic range and thereby make life more enjoyable. The cost to build such a high dynamic range receive converter is really not that much more than that of a conventional converter. If the dynamic range of the receive circuits is improved, transmitters can be evaluated more effectively. Who knows, you too may find a way to improve these circuits! (Is there any interest in designs for higher bands?)

acknowledgements

I would like to thank Dr. David Norton for his advice and reference material on the lossless feedback amplifier and Jim Reisert, AD1C, for his constructive comments and suggestions on this column. I'd also like to thank Jim Stitt, WA8ONQ, for his encouragement and comments on the performance of the converter circuits as they evolved.

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important VHF/UHF events in November, 1984

November 2:	*0330 UTC, predicted peak of Taurids meteor shower*
November 16:	*2100 UTC, predicted peak of Leonids meteor shower*
November 21:	*EME perigee*

ham radio

the ham notebook

the half-wave transmission line in bridge measurements

With the increasing availability of inexpensive, accurate RF impedance bridges, Amateurs are discovering how useful such a bridge can be in building and understanding antenna systems. However, unless bridge measurements are taken directly at the antenna, which is generally impractical, the effect of line length has to be taken into consideration when making calculations to determine the actual antenna impedance. This can be accomplished rigorously (by exact mathematical solution)[1] or in this case, simplified as in **eq. 1**. The transmission line is considered lossless, which for short lines is a valid assumption.

$$Z_{ANT} = \frac{Z_{in} - jZ_o \tan \phi}{Z_o - jZ_{in} \tan \phi} \bullet Z_o \quad (1)$$

where Z_{ANT} = antenna impedance
Z_{in} = transmission line impedance measured at bridge terminals
Z_o = characteristic transmission line impedance
ϕ = electrical length of transmission line in degrees

Now as everyone knows, regardless of the antenna or transmission line impedance, measurements made in multiple half-wavelengths repeat — i.e. the same value is seen regardless of whether measurements are taken directly at the antenna terminals or at an electrical half-wavelength along a transmission line. This is where I got into trouble. I measured the antenna impedance through a transmission line that was exactly one-half wavelength (180 degrees) at 3.75 MHz. Then I performed the same measurement (of the antenna) at 4.0 MHz *without* changing the transmission line length. How much difference could that make?

First off, my 180 degree half wavelength line at 3.75 MHz is actually *(4.0)(180)/3.75 = 192 degrees electrical length at 4.0 MHz*. And my measured impedance at the end of the line at 4 MHz was 175 + j 100 ohms impedance.

Substituting the values in **eq. 1**, I found:

$$Z_{ANT} = \frac{175 + j100 - (j)(50)(0.213)}{50 - (j)(175 + j100)(0.213)} \bullet (50)$$

$$Z_{ANT} = 70.8 + j\ 99.7 \text{ ohms impedance}$$

Well, my bridge had measured 175 + j 100 ohms, and my true calculated antenna impedance actually was 70.8 + j 99.7 ohms! That's really a large difference, and yet the electrical length of the line was only 12 degrees longer, or 6.7 percent longer.

So don't be fooled as I was into thinking that if you use an electrical half wavelength line at mid-band, your band end measurements will be close unless you actually correct for the few degrees as I first neglected to do.

Naturally, all antenna measurement calculations could have been done using a Smith chart, but to me the equation shows the impedance relationships involved more clearly.

Although the example was given for the 80-meter band, the same equation can be used for other bands, either for a band center half wavelength line, or as a general equation for use with any length of coax line, as long as you know the electrical length. And after I applied the technique just described, my previously measured data was much more meaningful.

reference

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William Vissers, K4KI

design superhet coilsets
with a microcomputer

Improve tracking with a one-minute interactive program

The design of superhet coilsets to ensure tracking for correct simultaneous tuning of preselector and oscillator circuits has taken on an undeserved air of mystery. Some receiver designers have avoided the problem by resorting to pre-peaked narrow-band circuits (not exactly ideal) or separate tuning of the preselector circuits, a throwback to the 1920s.

The project that led to this article was a receiver for 150-1560 kHz and 2.5-20 MHz in 6 bands with an IF of 1650 kHz. All the parts, including a zero-temperature coefficient (Invar) three-section tuning capacitor, were available. The specifications for the coilset — involving 18 coils, 18 trimmers, and 6 padders — had to be calculated. But how?

the one-minute solution

Years ago I found a set of formulae in the literature that seemed practical to use, although their derivation was not entirely clear to me. Using them, I wound and trimmed coilsets for my home-brew receivers. Because the receivers had worked well I hoped they would also track reasonably well. The problem was that calculating a single set took a whole day using a slide-rule and even after the advent of the pocket calculators, several hours. Half the time was spent in making mistakes and a quarter of the time in wondering if I had discovered all of them. I decided to write a design program based on the existing formulae to produce faster and more reliable results. Written in BASIC, the program reduced the chore to less than a minute!

Thereafter I added a subprogram for plotting the actual tracking curve on the screen. Such a curve shows the difference between the sum of signal and intermediate frequency on the one hand and the oscillator frequency on the other, for an entire tuning range. Ideally, it should be a straight line of zero error value. Without the aid of a computer, this would have taken me many hours to do. With the computer, the task took only a few seconds. As I watched, aghast, the errors ran off the screen as the program kept crashing for all types of designs.

After much thought, I concluded that the algorithms (published in a reliable journal) did not include the effect of stray capacitances such as coil winding capacitances. Even a few pF caused substantial tracking errors, up to 200 kHz or more.

An exhaustive search of the international literature led me to believe that tracking equations are avoided like the plague. Perhaps the subject is thought to be too boring and too difficult for Amateurs. I did find, however, two articles that were *not* over my head, but neither one could pass the computer test! A third article used complex math such as Vieta's theorem and

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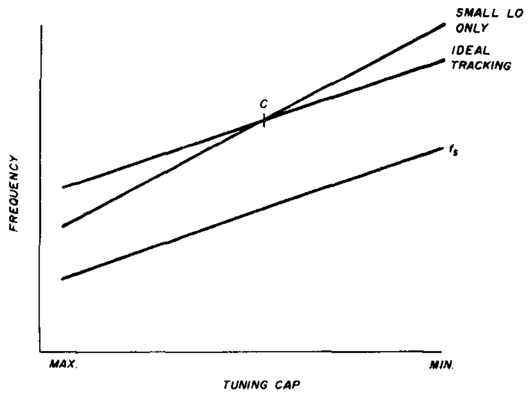


fig. 1. Why superhets require measures to ensure tracking.

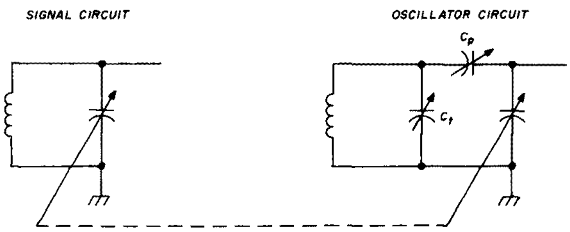


fig. 2. Runge's Remedy: trimmer and padder.

higher order polynomials, but the values given in the examples *did not check* with the formulae supplied! For instance, a trimmer of 46.36 pF in parallel with a tuning capacitor of 20-500 pF was said to give a tuning range of 1.5-5.0 MHz, whereas the actual value would be 1.5-4.3040 MHz. I almost gave up, but finally — in the absence of any information I could both understand and trust — I decided to try to work out the whole thing from the beginning. This is the result, but as you can imagine, I sometimes wonder how many published superhet designs really track.

the principle

In a superhet all signals (f_s) are mixed with a variable oscillator frequency (f_o) to produce a fixed intermediate frequency (f_i). The better solution is to choose an LO frequency higher than the incoming signals ($f_o = f_s + f_i$) as this reduces spurious responses due to oscillator harmonics. We will consider this case only; the alternative ($f_o < f_s$) is analagous with the proviso that the signal and oscillator coils are exchanged in the equations. The tuning capacitor will have identical sections; this is the reason tricks are necessary — with different sections for the tuning capacitor good tracking can be obtained, but such components are not readily available. The general equation for resonance is:

$$f^2 = \frac{25330}{L \cdot C} \text{ (MHz, pF and } \mu H) \qquad \textbf{(1A)}$$

and from this it is clear that tuning is not a linear function. If the oscillator coil (LO) were given less inductance than the signal coil L_s to obtain a high f_o, the rate of change of f_s and f_o would never yield a constant difference f_i. Fig. **1** shows the ideal tracking curve and the error resulting from using a smaller L_o only. If the receiver is made to track near the center of the tuning range (point C), f_o will be too high at the high end and too low at the low end. This could mean attenuation of the signal by 30 dB if you were using potted inductors in the 150-1600 kHz range.

A technique that solves this tracking problem, patented as far back as 1924 by W.T. Runge, is shown in **fig. 2**; a trimmer, C_t curtails the tuning capacitor and has the greatest effect at the high frequency end of the tuning range while the padder capacitor, C_p, does the same thing for the low end. This you know if you have ever aligned a receiver, but it is less widely known that adjustment of C_t and C_p only is not enough. **Fig. 3** shows a situation in which perfect tracking is obtained at both ends of the tuning range with nevertheless a substantial error in the middle, in this case because L_o is too small. Tracking, therefore, requires the determination and adjustment of C_t, L_o, and C_p at *three* frequencies and the curve will then approach a straight line with zero error at three points. Residual errors can be further reduced by shifting the outer two tracking points to the middle which will then result in a curve similar to that in fig. **4**.

The circuits shown in fig. **2** represent an idealized case; fig. **5** represents the actual situation including stray capacitances. As indicated by the vertical lines, a clear distinction should be made between the left half (all circuit elements associated with the coils) and the right half (all capacitances associated with the tuning capacitors and the receiver — for example, input capacitance of the active elements).

circuit elements

Both for practical and theoretical considerations all right half receiver sections must be identical. One important case to be considered is a Colpitts-type oscillator with capacitive divider, indicated by C_x and C_y in fig. **5**. An equivalent capacitor, C_e, is therefore added to the signal sections, where

$$C_e = (C_x \cdot C_y)/(C_x + C_y) \qquad \textbf{(2)}$$

The winding capacitance of the coils, C_w, is a problem because it is largely unpredictable (depending on

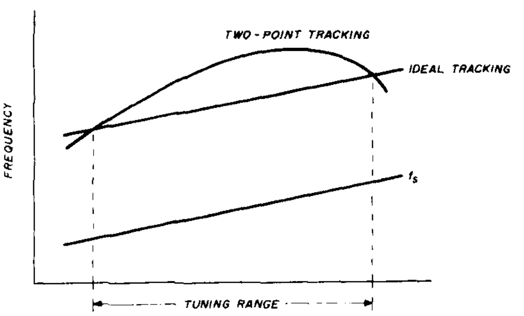


fig. 3. Two-point tracking is not enough!

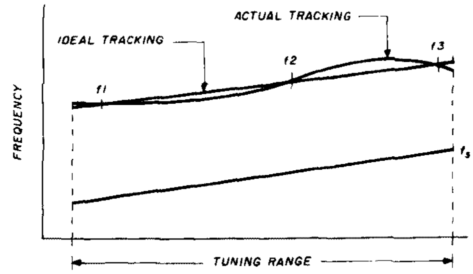


fig. 4. Three-point tracking.

winding technique) and is different for L_s and L_o. I solved the problem by combining C_w with C_t to form "coil capacitances" C_{cs} and C_{co}. The computer will ask you whether that value seems realistic; you can find out by simply winding a trial coil that resonates at the desired frequency with the tuning capacitor used. Then find its self-resonant frequency (without the tuning capacitor) with a dipper and from this derive its C_w. The same applies to the minimum and maximum capacitance of the tuning capacitor, C_{min} and C_{max}, and the wiring stray capacitance C_s. These are the known and unknown values of the circuit elements we shall work with.

calculating the values for the signal coils

These are easy to calculate. They are tuned by the combination on the right, which varies with tuning from a high to a low value:

$$C_H = C_{max} + C_s + C_e \qquad \textbf{(3)}$$

$$C_L = C_{min} + C_s + C_e \qquad \textbf{(4)}$$

Using the general resonance formula, it is seen that if a tuning range starts at f_L low frequency, an idealized coil without C_{cs} would tune to maximum frequency.

$$f_m = \sqrt{(C_H/C_L)} \cdot f_L \qquad \textbf{(5)}$$

Let the desired top frequency by f_H and let $R = f_H/f_L$ then:

$$R = \frac{25330/\sqrt{L_s \cdot (C_L + C_{cs})}}{25330/\sqrt{L_s \cdot (C_H + C_{cs})}} \qquad \textbf{(6A)}$$

Therefore: $R^2 = (C_H + C_{cs})/(C_L + C_{cs})$ **(6B)**

or: $C_H + C_{cs} = R^2 \cdot (C_L + C_{cs})$ **(6C)**

and from this it follows that the maximum allowed total capacitance across the signal coil (trimmer plus winding capacitance) can be

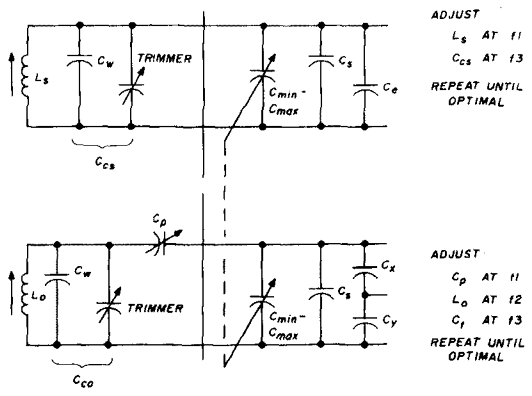


fig. 5. The actual tuned circuits.

$$C_{cs} = (R^2 \cdot C_L - C_H)/(1-R^2) \qquad \textbf{(6D)}$$

Allowing a reasonable value for the trimmer, you'll have to determine whether this leaves enough for the winding capacitance, by employing the procedure mentioned above — using a trial coil and dipper — if you're not sure.

example

Design a coil set for the medium wave band that tunes from 520-1620 kHz and incorporates a non-Colpitts oscillator. The tuning capacitor to be used provides from 15-500 pF capacitance and the stray capacitance is 15 pF. C_{cs} when evaluated turns out to be 25.7 pF which is a low value for this range. The medium wave band, with a frequency ratio of over 1:3 requires an approximate 1:10 capacitance ratio and this leaves little leeway for strays.

The value of the signal coil is:

$$L_s = 25330/f_L^2 \cdot C_H \qquad \textbf{(1B)}$$

and in this example $L_s = 173.25\ \mu H$

The signal circuits are now completely defined; additional preselector circuits must be identical.

determining the tracking points

One tracking point must be at the center of the range. This can be found by taking the geometric means of the two end frequencies: $\sqrt{f_L \cdot f_H}$. The outer tracking points must be shifted slightly toward the center. The amount will vary with the tuning range. Empirically, the following factor gives good results:

$$Q = 1 + (f_H/40 \cdot f_L) \quad \textbf{(7)}$$

You can easily change this factor in the program and observe the result on the display. The greatest deviation will occur at the high end, which is acceptable because the bandwidth of the signal coils is also a maximum there. The tracking (and trim-) points are:

$$f_1 = f_L \cdot Q \text{ (low end)} \quad \textbf{(8)}$$

$$f_2 = \sqrt{f_L \cdot f_H} \text{ (center)} \quad \textbf{(9)}$$

$$f_3 = f_H/Q \text{ (top end)} \quad \textbf{(10)}$$

which calculates to be 560, 918, and 1503 kHz, respectively in the example outlined earlier. As the operator tunes to these frequencies, he will set the tuning capacitor to values that can be calculated from the resonance formula after subtracting C_{cs}. These values (for the right half of the signal sections) are:

$$C1 = \frac{25330}{f_1^2 \cdot L_s} - C_{cs} \quad \textbf{(11)}$$

$$C2 = \frac{25330}{f_2^2 \cdot L_s} - C_{cs} \quad \textbf{(12)}$$

$$C3 = \frac{25330}{f_3^2 \cdot L_s} - C_{cs} \quad \textbf{(13)}$$

In the oscillator section, the total capacitance to the right will be the same as above (C_s and C_e are identical), so with the three capacitances $C1$, $C2$, and $C3$ the oscillator must tune to $f_1 + f_i$, $f_2 + f_i$ and $f_3 + f_i$.

determining the oscillator circuit elements

Taking into account the total coil capacitance for the oscillator circuit and the padder, gives the following three equations:

$$f_1 + f_i = 25330/\sqrt{C_{co} + \frac{C_p \cdot C_1}{C_p + C_1} \cdot L_o} \quad \textbf{(14)}$$

$$f_2 + f_i = 25330/\sqrt{C_{co} + \frac{C_p \cdot C_2}{C_p + C_2} \cdot L_o} \quad \textbf{(15)}$$

$$f_3 + f_i = 25330/\sqrt{C_{co} + \frac{C_p \cdot C_3}{C_p + C_3} \cdot L_o} \quad \textbf{(16)}$$

The solution to these equations means that the tracking error is indeed made equal to zero at these points.

Let's tackle the padder first. Define the ratio $(f_1 + f_i)/(f_3 + f_i)$ as "A" and call $(f_2 + f_i)/(f_3 + f_i)$ as "B," then by dividing eq. **14**/eq. **16** we find that:

$$C_{co} = \frac{A^2 \cdot C_p \cdot C_1/(C_p + C_1) - (C_p \cdot C_3)/(C_p + C_3)}{1 - A^2} \quad \textbf{(17)}$$

Similarly, division of eq. **15** by eq. **16** yields:

$$C_{co} = \frac{B^2 \cdot C_p \cdot C_2/(C_p + C_2) - (C_p \cdot C_3)/(C_p + C_3)}{1 - B^2} \quad \textbf{18)}$$

These two equations can be used to solve for C_p:

$$C_p = \frac{C_1C_2(B^2 - A^2) + C_1C_3(A^2B^2 - B^2) + C_2C_3(A^2 - A^2B^2)}{C_1(A^2 - A^2B^2) + C_2(A^2B^2 - B^2) + C_3(B^2 - A^2)}$$

(This looks neater when the terms in parentheses are called X, Y, and Z respectively, as in the program.) C_{co} is found by entering C_p in eq. **17** or **18** and then:

$$L_o = \frac{25330}{(f_2 + f_i)^2 \cdot \left(C_{co} + \frac{C_p \cdot C_2}{C_p + C_2}\right)} \quad \textbf{(19)}$$

This completes the coilset design. In this example: $C_p = 585.7\ pF$, $C_{co} = 41.9\ pF$, $L_o = 84.66\ \mu H$ for $f_i = 450\ kHz$.

tracking curve

Plotting the tracking curve on the screen depends on the graphic capability of your microcomputer. High resolution plotting in assembly language is definitely not necessary: what you want to see is the general trend and the peak errors, not an accurate graph. The method used in plotting the tracking curve is described below.

First the tuning range is divided into as many equal sections as the micro has columns. Because of rounding off, the range must be extended above f_H; otherwise, the plot sometimes won't reach this value. For a 40-column machine and half a column extra margin the command would be:

for f = f_L to f_H + (f_H − f_L)/80 step (f_H − f_L)/40

For each f calculate the total capacitance across the signal coil:

$$C = 25330/(L_s \cdot f^2) \quad \textbf{(1C)}$$

So the capacitance to the right of the lines is equal to $C_v = C - C_{cs}$. From this we can calculate the oscillator frequency correspnding to f:

$$f_o = \sqrt{\frac{25330}{L_o\left(C_{co} + \frac{C_p \cdot C_v}{C_p + C_v}\right)}} \quad \textbf{(20)}$$

Then print $f_o - f - f_i$ together with f for a tracking table or use $f_o - f - f_i$ as the row- and f as the

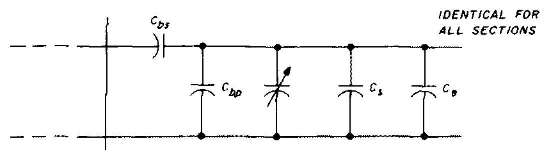


fig. 6. Series and parallel bandspreading.

column parameter. To avoid crashing from incorrect row data, first display the table and examine the size of the error values, then enter a corrective multiplication factor via an INPUT command if necessary.

For this example the errors (rounded off) were −3.2, +4.26, −6.97 and +8.5 kHz. This indicates excellent tracking and implies that everything is correct — an assurance I never had before.

the program

The program listing in **table 1** was written for the TI-99/4A which is suitable for this application because of its upper and lower case capability, which makes for easy reading, its wide choice of variables, and high degree of precision in mathematics. Note that when the program runs, very small values are subtracted from other very small values so that you may have to use double precision. I have left out everything that is not essential and used the simplest BASIC commands for easy translation. One version I have draws all coils and capacitors in HiRes and inserts their values next to them for added finese. The listing takes less than 4k of memory, and can still be adapted for your particular machine. A printer is not necessary; the program warns you to "NOTE" the values displayed.

bandspreading

As supplied, the program already includes parallel bandspreading, because parallel capacitance is added in the form of C_{cs} by your independent choice of f_H. But let's look at bandspreading in more detail (**fig. 6**).

Parallel bandspreading with C_{bp} improves the tuning curve. Ordinary tuning capacitors provide logarithmic coverage, with compression at the high end. For straight-line frequency coverage you need a tuning capacitor with pointed plates (the BC-221 has one), but these are exceptionally rare. Adding C_{bp} reduces the top end compression. The disadvantage of adding parallel capacitance is that the L/C ratio of the tuned circuit is lowered and Q is reduced, especially at the high end. You may assume that the effect on the Q is acceptable if C_{bp} is smaller than:

$2 \cdot \{C_L$ + *reasonable value for trimmer (e.g. 15 pF)* + *value found for* $C_w\}$

Series bandspreading C_{bs} raises the L/C ratio, but

tabie 1. Program listing for superhet coilset design.

```
100 CALL CLEAR
110 PRINT "  SUPERHETERODYNE COILSETS": : : : : : :"   for Ham Radio Magazine":
: :"by Frithjof A.S. Sterrenburg": : : :"          ** 1984 **": : : : : : :
120 FOR DELAY=1 TO 1000
130 NEXT DELAY
140 CALL CLEAR
145 REM for colpitts type oscillator
150 INPUT "does the oscillator have a  capacitive divider? y/n":A$
160 IF A$="y" THEN 170 ELSE 230
170 INPUT "enter Cx,Cy ":CX,CY
180 CE=(CX*CY)/(CX+CY)
190 CALL CLEAR
200 PRINT "an extra parallel condenser Ce of";CE,"pF will be specified for thesi
gnal sections in the finalprint-out"
210 FOR DELAY=1 TO 1000
220 NEXT DELAY
230 CALL CLEAR
235 REM define tuning range
240 INPUT "enter minimum capacitance oftuning condenser (pF) ":CMI
250 PRINT : :
260 INPUT "enter maximum capacitance oftuning condenser (pF) ":CMA
270 PRINT : :
280 INPUT "enter estimated stray capa- citance in pF ":CS
290 PRINT :"Ce calculated earlier will  be printed at the end if it was required
": :
300 CL=CMI+CS+CE
310 CH=CMA+CS+CE
320 INPUT "enter lower frequency of tu-ning range in MHz ":FL
330 PRINT : :
340 FM=SQR(CH/CL)*FL
350 PRINT "the tuning range would be";FL;"to";FM;"MHz without trimmer or wind-in
g capacitance": :
360 INPUT "enter desired top frequency of tuning range ":FH
370 R=FH/FL
380 CCS=(CH-(R^2*CL))/(R^2-1)
390 CALL CLEAR
400 PRINT "                 NOTE:": :"total desired coil capacitor(trimmer + coil capa
citance)=";CCS;"pF": :
410 INPUT "does this seem feasible? y/n ":B$
420 IF B$="n" THEN 360
430 CALL CLEAR
440 CL=CL+CCS
450 CH=CH+CCS
460 LS=25330/(FL^2*(CH))
465 REM total specification of signal coils
470 PRINT "                 NOTE:": :"Ls=";LS;" microH": :"assumed stray capacitance=
";CS;"pF": :"trimmer + coil capacitance= ";CCS;"pF": :
480 PRINT "for all preselector sectionsadd extra parallel capacitorof";CE;"pF(Ce
)": :
490 INPUT "enter c to continue ":C$
500 IF C$="c" THEN 510 ELSE 490
510 Q=1+(FH/FL)*0,025
520 F1=FL*Q
530 F2=SQR(FL*FH)
540 F3=FH/Q
550 CALL CLEAR
555 REM determine the tracking points and oscillator components
560 PRINT "                 NOTE:": :"low trimpoint=";F1;" MHz":"mid trimpoint=";F2;"
MHz":"high trimpoint=";F3;" MHz": : : :
570 C1=(25330/(F1^2*LS))-CCS
580 C2=(25330/(F2^2*LS))-CCS
590 C3=(25330/(F3^2*LS))-CCS
600 INPUT "enter intermediate frequencyin MHz ":FI
610 LET A=((F1+FI)/(F3+FI))
620 LET B=((F2+FI)/(F3+FI))
630 X=B^2-A^2
640 Y=(A^2*B^2)-B^2
650 Z=A^2-(A^2*B^2)
660 CP=((X*C1*C2)+(Y*C1*C3)+(Z*C2*C3))/((Z*C1)+(X*C3)+(Y*C2))
670 CCO=(A^2*(CP*C1)/(CP+C1)-(CP*C3)/(CP+C3))/(1-A^2)
680 LO=25330/((F2+FI)^2*(CCO+((CP*C2)/(CP+C2))))
690 CALL CLEAR
695 REM total specification oscillator components
700 PRINT "                 NOTE:": :"padder=";CP;"pF": :"trimmer including winding
capacitance=";CCO;"pF":

710 PRINT "(for signal coil";CCS;"pF was allowed)": :"Lo=";LO;"microH": :"capaci
tive divider";CX;"and";CY;"pF": :
720 INPUT "enter 1 for next set,2 for  tracking table,3 for graph, 4 to end ":K
730 ON K GOTO 140,740,820,990
740 CALL CLEAR
745 REM   printout of tracking errors for range,for 30 frequencies
750 FOR F=FL TO FH+(FH-FL)/60 STEP (FH-FL)/30
760 C=25330/(LS*F^2)
770 CV=C-CCS
780 FO=SQR(25330/(LO*(CCO+((CP*CV)/(CP+CV)))))
790 PRINT F*1000,(F+FI-FO)*1000
800 NEXT F
810 GOTO 720
820 CALL CLEAR
825 REM plot of tracking curve for 24 lines/30 columns
830 INPUT "for error plot enter multi- plication factor.maximum de-viation=10 ":
M
840 CALL CLEAR
845 REM   draw plotting grid (for 99/4A)
850 CALL CHAR(128,"10101010FF101010")
860 PRINT TAB(16);"factor=";M: :
870 CALL HCHAR(13,1,128,31)
880 CALL VCHAR(1,16,128,24)
890 Y=1
895 REM  for 30-column machine this is identical to 750-780 inclusive and can b
e a gosub
900 FOR F=FL TO FH+(FH-FL)/60 STEP (FH-FL)/30
910 C=25330/(LS*F^2)
920 CV=C-CCS
930 FO=SQR(25330/(LO*(CCO+((CP*CV)/(CP+CV)))))
940 X=(F+FI-FO)*1000*M
945 REM plot asterisks
950 CALL HCHAR(12+X,Y,42)
960 Y=Y+1
970 NEXT F
980 GOTO 720
990 END
```

tuning at the top end is even more compressed. Series and parallel bandspread techniques can, of course, be combined. For series bandspreading, a series capacitor C_{bs} is added to the right hand side and this is identical for all tuning capacitor sections. In fact, we construct a new tuning capacitor with different C_{min} and C_{max} and then run the program as normally.

To determine C_{bs} for a desired range we define new constants. Let $C_{cs} = j$, desired range $k = f_H/f_L$, $m = C_H$, $n = C_L$ and the unknown bandspread capacitor = b. Then starting with the equation for the tuning range:

$$k^2 = \frac{j + \dfrac{b \cdot m}{b + m}}{j + \dfrac{b \cdot n}{b + n}} \qquad \textbf{(21A)}$$

you will arrive at this:

$$(b \cdot m) \cdot (b + n) - (b + m) \cdot [\{k^2 \cdot n + (k^2 - 1) \cdot j\} \cdot b + (k^2 - 1) \cdot j \cdot n] = 0 \qquad \textbf{(21B)}$$

Now let's call $k^2 \cdot n + (k^2 - 1) \cdot j = p$ and $(k^2 - 1) \cdot j \cdot n = q$, then:

$$b^2 \cdot m + b \cdot m \cdot n - (q \cdot b^2 + p \cdot b + q \cdot m \cdot b + p \cdot m) = 0 \qquad \textbf{(21C)}$$

$$\text{or } b^2 + \frac{m \cdot n - p - q \cdot m}{(m - q)} \cdot b - \frac{p \cdot m}{(m - q)} = 0 \qquad \textbf{(21D)}$$

If we call the two fractions "r" and "s" respectively, this is the common equation $b^2 + r \cdot b - s = o$. One root is negative, the other is the value of the bandspread capacitor:

$$C_{bs} = \frac{-r + \sqrt{r^2 + 4 \cdot s}}{2} \qquad \textbf{(22)}$$

Although all bandspread capacitors have been lumped with the right hand half of the circuits in this derivation, they are switched for other ranges and will therefore be *physically present in the left hand half* (in the coil cans, for instance). In the end, you can properly combine all the capacitances calculated.

conclusion

Although any coilset can now be designed in a few minutes, the main value of this project was something else. For the first time, I could plot tracking curves, and these showed that all the literature I could lay hands on gave unreliable formulas! The microcomputer freed me from the nightmare of calculating, so I could begin to *think* and rediscover a bit of neglected theory for myself. A microcomputer may have its practical uses, but above all, it's a powerful tool for learning!

ham radio

receiver sweep alignment system

No sweep generator?
Try these handy
throwaway circuits

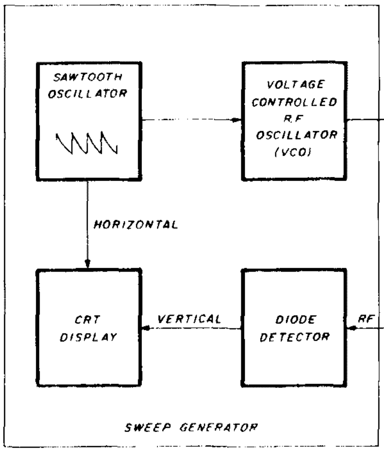


fig. 1. Sweep generator; the CRT (Cathode Ray Tube) displays relative amplitude of the swept filter output.

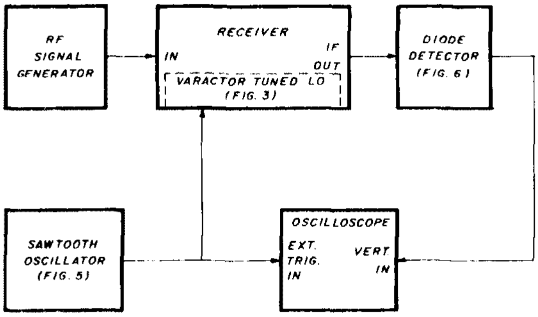


fig. 2. Minimal configuration sweep system uses the receiver LO (Local Oscillator) as the VCO (Voltage Controlled Oscillator).

When a circuit just "doesn't sound right," the obvious solution is to use a sweep generator to evaluate and align the filters. But a sweep generator may not be available; for most of us, it's too expensive a piece of equipment for the occasional use it receives.

This article describes how I approached this problem in the development of an SSB receiver by adding a few extra circuits during construction, then removing and discarding them after use. A separate signal generator and oscilloscope were also required.

sweep measurement basics

Fig. 1 illustrates a typical sweep measurement system. The sawtooth oscillator generates a voltage "ramp" which tunes the voltage controlled oscillator frequency across the filter passband. For small frequency changes, the voltage controlled oscillator often uses a varactor ("varicap") diode to change circuit capacitance. The amplitude of the RF/IF signal coming through the filter varies with (and helps define — **Ed.**) the filter's frequency response, and is detected by the diode detector. The detector output is displayed by the oscilloscope vertical channel while the sawtooth oscillator drives the oscilloscope horizontal channel in step. The resulting display plots the filter's amplitude versus frequency response. This display is used to align filters since it gives an instant indication of circuit adjustment results. This is very handy when a large number of interrelated circuit manipulations must be made.

The proposed minimal sweep alignment system uses the existing receiver local oscillator as the voltage controlled oscillator. This is similar to a panoramic receiver with a much smaller sweep range. **Fig. 2** shows this scheme as implemented in an HF SSB receiver project. Three new circuits are added; a varactor tuning diode, a diode AM detector, and a sawtooth oscillator circuit.

detailed circuit description

Because the three circuits were designed to be dis-

By Cliff Klinert, WB6BIH, 1126 Division Street, National City, California 92050

carded after use, careful consideration was given to parts availability.

The familiar 1N4000-series of silicon rectifiers make good "varactor" diodes when biased in the linear region. **Fig. 3** shows the varactor tuning circuit using a 1N4007 connected to a typical 5 MHz oscillator tank circuit. The circuit was tested by applying an adjustable DC bias voltage and measuring the corresponding frequency with a counter. **Fig. 4** shows the results of this experiment for DC bias voltages from 14 to almost 28 volts. Note that the curve is almost linear. Good linearity throughout the sweep system is required to provide an undistorted picture of the filter's passband response.

The sawtooth oscillator, the heart of the system, generates a periodic linearly increasing voltage ramp waveform. If the oscilloscope used in this project had provided a sweep output connection, the external oscillator might not have been required. The varactor circuit requires a linear sawtooth voltage providing a 14 to 24 volt ramp per frequency sweep. Simple sawtooth

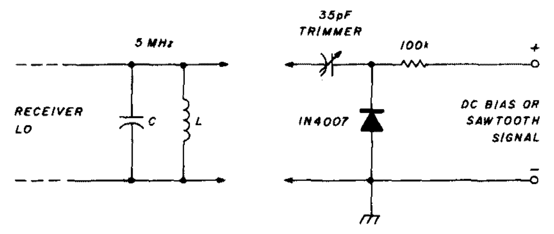


fig. 3. The varactor diode acting as a voltage variable capacitor changes the resonant frequency of an LC circuit in step with a varying DC bias.

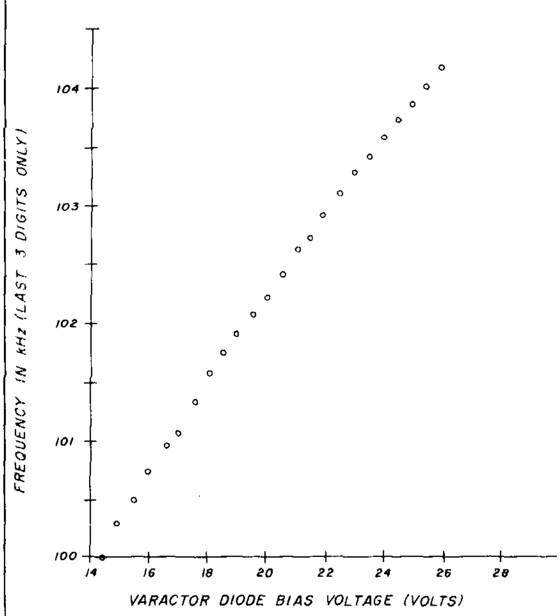


fig. 4. Varactor diode tuning curve is a plot of actual data taken from the circuit of fig. 3. The range of the swept frequency is adequate for the SSB filter tested.

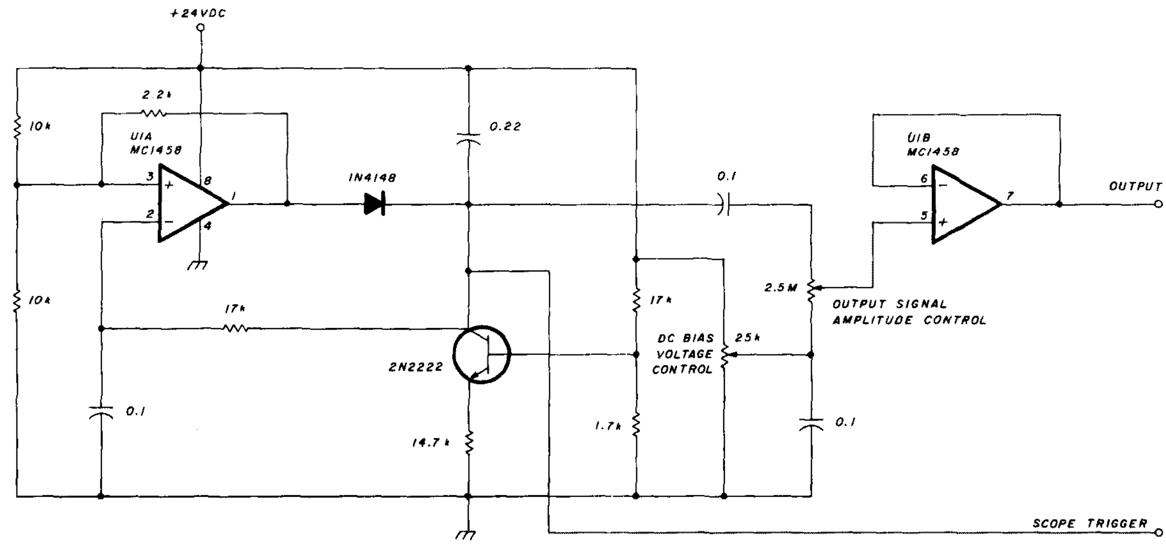


fig. 5. The sawtooth oscillator uses an inexpensive IC (MC1458) to both generate the ramp waveform and buffer its output. Mylar capacitors are used for best results.

oscillator circuits are normally designed around unijunction transistors and in this case, to keep costs down, a less expensive IC, an MC1458, (U1 of **fig. 5**), was used. The 0.22 μF capacitor is charged by a 2N2222 constant current source until the MC1458 "triggers" and briefly shorts the capacitor back to the positive supply voltage line. The cycle is then immediately repeated. Since the 2N2222 collector current is determined by its base and emitter bias circuit, it is nearly constant while charging the 0.22 μF capacitor. This constant current ensures a linear capacitor voltage rise with time. The other half of the MC1458, U1B, is used as a voltage follower and provides a low impedance, higher current version of the sawtooth voltage at its output. The 2.5 Megohm potentiometer controls the amount of frequency excursion, while the 25 kilohm potentiometer tunes the voltage controlled oscillator frequency.

The diode detector provides a DC voltage that is proportional to the RF swept signal output amplitude. A hot carrier diode such as a 1N5711, shown in **fig. 6A**, can be used. If a hot carrier diode is not available, an inexpensive germanium diode that is slightly forward biased can be used in its place as shown in **fig. 6B**. A DC return is required for the *biased* diode, and the diode impedance decreases as the current in-

fig. 7. Oscilloscope photographs. Unless otherwise noted, the sweep speed is 5 milliseconds per division and vertical sensitivity is 0.1 volt per division. The unit under test is a 9.0 MHz SSB IF amplifier using an MC1350 and surplus crystal filter.

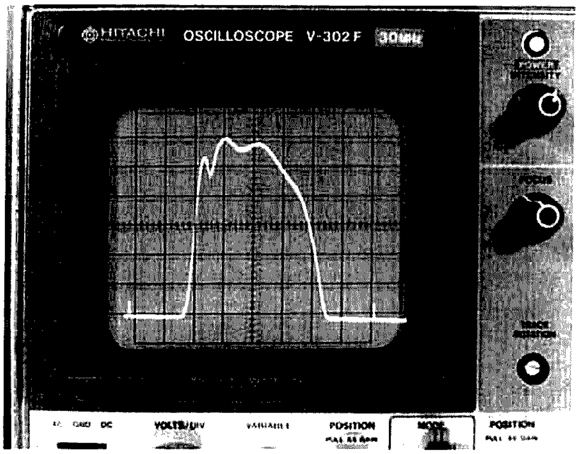


fig. 7A. This shows a typical frequency sweep about 3.5 kHz wide.

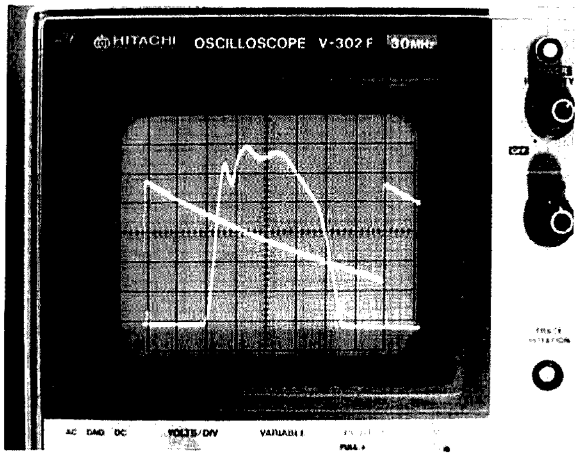


fig. 7B. The second channel is used to display the sawtooth signal showing its relationship to the frequency sweep.

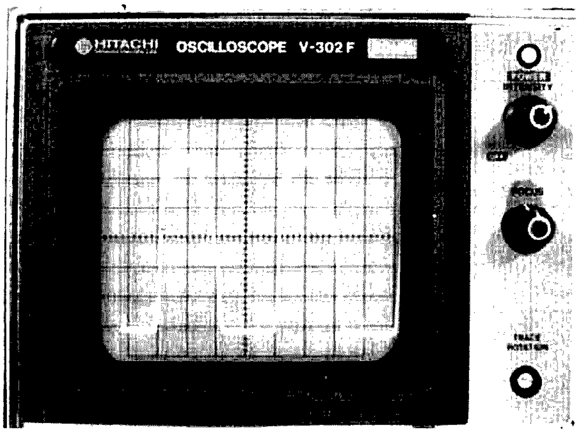


fig. 7C. The sweep speed is too low (10 milliseconds per division). This indicates that the speed should be adjusted to provide the desired display.

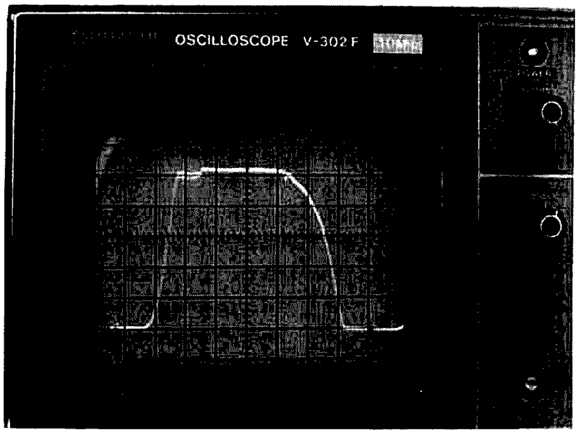


fig. 7D. The IF amplifier is "flat topping." The gain or input signal must be reduced to remove the distortion at the top of the display.

creases. A high impedance tuned circuit will be "shorted out" by the biased diode. Adjusting the diode bias control will produce a sharp peak in the detected signal output. Reducing the control's resistance near zero will, of course, destroy the diode. A silicon diode such as a 1N914 works very poorly as a detector.

construction and installation proceed smoothly

The circuit requires 24 volts DC which can be provided by either a bench power supply or batteries. Connect the varactor circuit (**fig. 3**) to the receiver local oscillator. Apply a variable DC bias voltage and adjust the varactor-tuned circuit to achieve results similar to those in **fig. 4**. Make a notation of the voltage variation required to obtain the full frequency sweep.

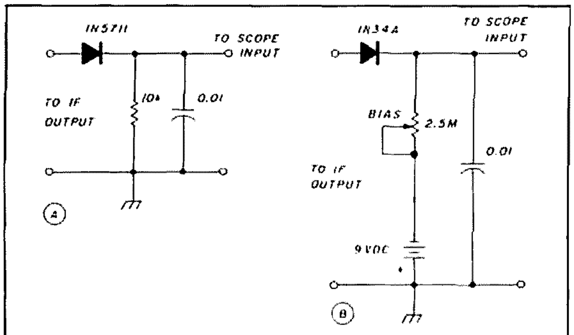


fig. 6. The diode detector is connected between the IF or filter output coupling link and the oscilloscope vertical input. The detector can also be used at the audio output.

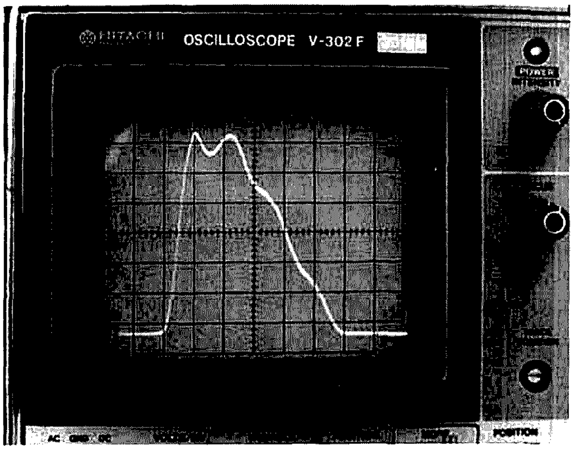


fig. 7E. Crystal filter load impedance is too low, resulting in a noticeable loss of audio frequency response and confusion in setting the BFO frequency.

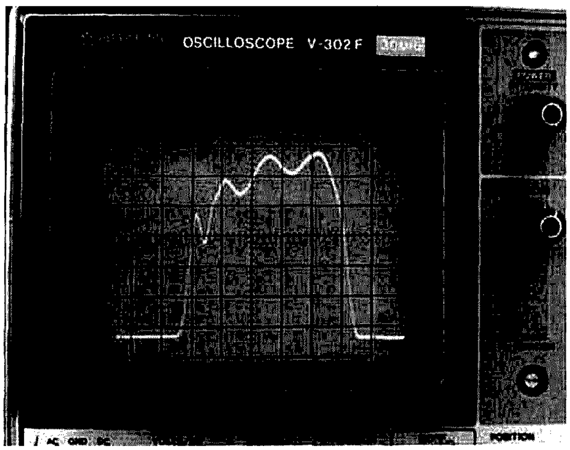


fig. 7F. The crystal filter load impedance is high, about 6000 ohms.

Assemble the sawtooth oscillator (**fig. 5**) and check the results with the oscilloscope. The output voltage should be capable of a swing nearly equal to the power supply limits with a period of approximately 50 milliseconds. The circuit can be assembled on a circuit board, but for temporary use, just solder the components together by their leads on the bench. Stray noise pickup may be a problem, but the large signal makes this unlikely. The varactor diode circuit is quite sensitive because of its high impedance, but no problems were encountered when connected to the sawtooth oscillator output.

Connect the detector probe (**fig. 6**) to the receiver IF amplifier output at a point of maximum available signal. Some experimentation is necessary with a signal generator and oscilloscope to obtain the maximum detected output without saturating the IF amplifier.

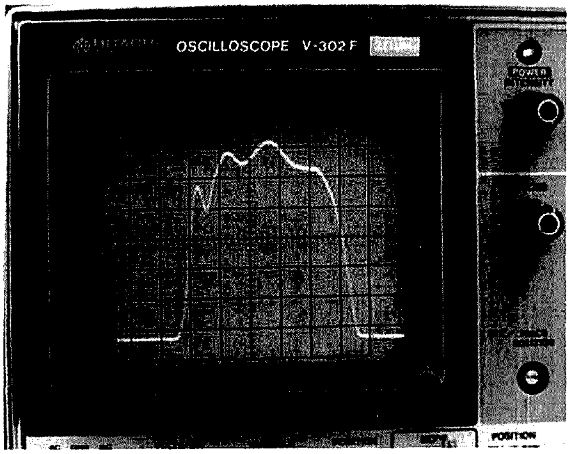


fig. 7G. The best response was obtained with approximately 3000 ohms load impedance. This is an unusually high load impedance for a crystal filter, and may indicate some problem in the filter.

The oscilloscope must be DC coupled with at least 0.1 volt per cm sensitivity.

Connect the sawtooth oscillator output to the varactor circuit. It's easier to start with a relatively low sweep amplitude when finding the frequency, so reduce the sweep signal amplitude in the beginning. Connect the diode detector output to the oscilloscope vertical input and synchronize the oscilloscope sweep from the sawtooth oscillator signal. The sawtooth signal could also be connected to the oscilloscope horizontal amplifier input if access is available. Tune the receiver to the signal generator frequency and make adjustments as required. Slowly increase the sawtooth signal amplitude until a sweep display indication is obtained. **Fig. 7** shows the results obtained from an SSB crystal filter sweep. In this case, the crystal filter output impedance was varied with a potentiometer in series with the filter output. The photographs show IF sweeps, but similar results were obtained by connecting the detector to the outputs of the product detector and audio amplifier. This provides analysis of other points in the receiving system that would be useful for troubleshooting or design evaluation.

conclusions

This article presented the concept of using expendable circuits as built-in test equipment for use during project construction. A handy sawtooth oscillator was presented for those who collect simple circuits for afternoon projects. This oscillator will find many applications in oscilloscope or spectrum analysis projects. No construction details were presented, since the concept was to show that circuits can be assembled without circuit boards for prototype or temporary use and discarded later.

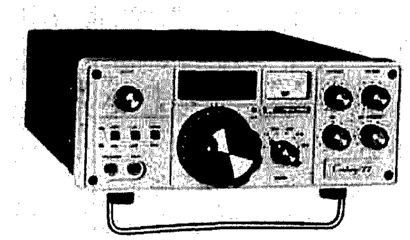
ham radio

NEW products

the Century 22

Ten-Tec has announced the return of the Ten-Tec Century *transceiver*. The Century 22 is a 50-watt, 6-band CW transceiver that features a variable audio filter, automatic gain control, an SWR bridge, automatic level control, and an electronically switched "S" meter.

The Century 22 measures 4 × 10 × 10.5 inches (25 × 101 × 29 cm), weighs 6 pounds (2.7 kg), and is priced at $389.

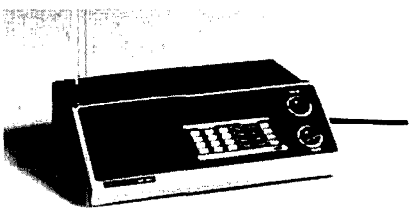


For information, contact Ten-Tec, Inc., Sevierville, Tennessee 37862.

7-band scanning radio

Heath Company has introduced the only kit-built scanner to cover aircraft, marine and public service bands, *all in one unit*. The GR-740 40-Channel Scanning Radio covers all seven UHF/VHF radio bands, scans 40 user-selected frequencies and provides direct access to any frequency in the seven bands.

The 24-key keyboard is divided into program and operate sections for simplified operation. Forty different channels (frequencies) are easily programmed into the two 20-channel memory banks. Either bank can be scanned at five or 15 channels per second; the GR-740's search can be programmed or changed at the touch of a button. A priority channel can be sampled every two seconds, with interruption when a signal is detected.



Patented track tuning permits receiving frequencies across the full band without adjustments; circuitry is automatically aligned to each monitored frequency. A large digital, front-panel display shows the channels and features selected. All circuit boards are factory-assembled and pre-aligned to ensure that even the first-time kit builder can build and operate one of the world's best scanning radios, with a minimum of time and at a substantial savings.

For more information about the GR-740 40-Channel Scanning Radio, contact Heath Company, Dept. 150-315, Benton Harbor, Michigan 49022.

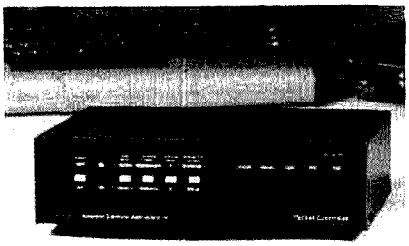
Circle #301 on Reader Service Card.

TAPR packet radio controller

Advanced Electronic Applications, Inc. has announced the introduction of the Model PKT-1, packet radio controller, through an arrangement with Tucson Amateur Packet Radio, Inc. (TAPR), Tucson, Arizona. While the end user price is $589.95, Amateur Radio operators can take advantage of a discounted price of $499.95 through participating AEA dealers.

The PKT-1 is a packaged and warranted version of the well-known do-it-yourself TAPR kit board with version 3.1 software. The purchase price includes application assistance and a year's conditional warranty.

Packet Radio is a burst mode of data or text transmission utilizing AFSK, FSK, or PSK modulation. On VHF it runs at 1200 Baud typically and uses CRC error checking, ensuring an extremely low error rate. Multiple users may share a simplex or duplex channel simultaneously on a timeshare *multiplexed* basis.



Any packet station using the PKT-1 may operate as a store-and-forward repeater (Digipeater) for someone else's transmission while concurrently functioning as a regular packet station. Up to 8 Digipeating stations may be used between two terminal stations. Digipeating allows routing the transmission path around physical obstacles blocking a line-of-sight radio path and allows extending the link beyond line-of-sight distances.

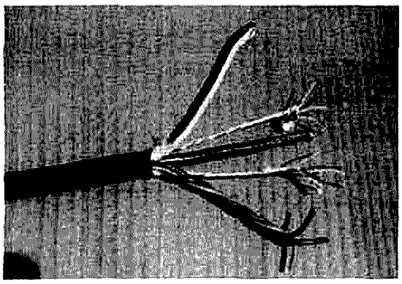
For detailed information, contact Advanced Electronic Applications, Inc., P.O. Box C2160, Lynnwood, Washington 98036-0918

Circle #304 on Reader Service Card.

NEW products

TVRO cable

Nemal Electronics International of North Miami, Florida, has introduced a new addition to its line of direct burial combination cable for use in TVRO installations. Nemal type-4 satellite control cable is the first combination cable available to the satellite industry containing an RG-6/U, 18 gauge, 95 percent copper shielded signal cable. SCC-4 also contains two conductors of 12 gauge, three conductors of 18 gauge, three conductors of 20 gauge shielded plus drain wire, and three conductors of 22 gauge shielded plus drain wire.



All Nemal satellite control cables utilize a patented direct-burial polyethylene jacket as well as tinned copper drain wires. Nemal also offers a complete line of over 500 types of cable, connectors, and SMATV products.

For additional information, contact Nemal Electronics International, Inc., 12240 N.E. 14th Avenue, North Miami, Florida 33161.

Circle #303 on Reader Service Card.

outdoor scanner antenna

Hamtronics, Inc. has announced a new antenna for scanner and monitor buffs. The compact ACT-1 Power Antenna, which may be installed easily on the side of a house, outside a window, in an attic, etc., without any special masts or brackets, is a broadband whip antenna with a low-noise preamplifier in its base. Although smaller than a full-size outdoor antenna (only 25 inches tall), the ACT-1 provides good coverage of distant signals and often outperforms larger antennas because of its active booster amplifier. A low-noise microwave transistor in the preamp provides excellent results from 30 MHz right up through the new 800-MHz band, and covers low-band, high-band, and UHF.

The ACT-1 Power Antenna is mounted to any flat vertical surface with four wood screws. The 50-foot cable plugs directly into the "antenna" and "12V" jacks on the rear of most scanner radios. If your particular scanner doesn't have

NEW products

a 12V terminal, a simple 12 VDC plug-in adapter is available.

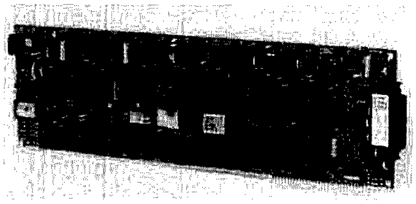
The price of the ACT-1 Power Antenna is only $79 plus $3 for shipping and handling.

For more information, contact Hamtronics, Inc., 65 Moul Road, Hilton, New York 14468-9535.

Circle #305 on Reader Service Card.

RTTY interface

HAL Communications RTTY Personal Computer Interface, PCI-2000, is a real RTTY modulator/demodulator, not a "computer compromise." All three shifts (170-425-850) transmit and receive. The full "103 type" modem for up to 300 baud may be set for either FDX (answer or originate) or HDX (either set on tones). The "202 type" modem may be used HDX up to 1200 baud, and may be jumper-selected for either "COM1" or "COM2" operation. Compatible with existing PC communications software (external DAA required for phone line connections, the PCI-2000 offer the multimode features of the CT2100 and CT2200 plus companion software that includes features such as split screen, TX and RX buffers, HERE IS storage, and disk message storage and retrieval. Of course, all PCI-2000 modes are set with the Personal Computer's FN keys.



For further information, contact HAL Communications Corporation, P.O. Box 365, Urbana, Illinois 61801.

Circle #302 on Reader Service Card.

150-MHz mini-catalog

Sinclair Radio Laboratories has issued a new mini-catalog describing its line of 150-MHz products, which includes base station antennas, transmitter combiners, duplexers, receiver multicouplers and ferrite isolators.

Featured in this line-up is the Q-Circuit Base Station Duplexer, Model Q-201G, a six-cavity unit that provides high attenuation at close frequency separations in the 132-174 MHz band. Its Q-Circuit design provides 100 dB isolation at 300 kHz spacing with 50 dB mid-band isolation.

For a copy, contact Sinclair Radio Laboratories, 122 Rayette Road, Concord, Ontario, Canada L4K 2G3.

Circle #306 on Reader Service Card.

✓ 208

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DEADLINE 15th of second preceding month.

SEND MATERIAL TO: Flea Market, Ham Radio, Greenville, N. H. 03048.

OSLs & RUBBER STAMPS — Top Quality! Card Samples and Stamp info — 50¢ — Ebbert Graphics 5R, Box 70, Westerville, Ohio 43081.

IBM-PC RTTY ASCII/BAUDOT/CW send and receive. Split screen, buffers and features beaucoup. SASE to: E. Alline, 773 Rosa, Metairie, LA 70005.

CX7 REPAIRS. 415-549-9210.

WANTED: General Radio 874MR mixer. Rudolf Six, KA8OBL, 30725 Tennessee, Roseville, MI 48066.

TRAVEL-PAK QSL KIT — Converts post cards, photos to QSLs. Stamp brings circular. Samco, Box 203-c, Wynantskill, New York 12198.

WANT old antenna books, handbooks, CQ, pre-1940 OST. Tad Cook, KT7H, 5519 — 12th N.E., Seattle, WA 98105.

CUSTOM MADE embroidered patches. Any size, shape, colors. Five patch minimum. Free sample, prices and ordering information. Hein Specialties, Inc., Dept. 301, 4202 N. Drake, Chicago, IL 60618.

QST, 1959-present, $150. HR, complete, $150. RF Design, complete except V2N1, $40. you ship. Books, SASE Meyer, 400-1736 N. River Rd., W. Lafayette, IN 47906.

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Holland

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I-20132 Milano
Italy

Ham Radio Switzerland
Karin Ueber
Postfach 2454
D-7850 Loerrach
West Germany

Ham Radio England
c/o R.S.G.B.
Alma House
Cranborne Road
Potters Bar
Herts EN6 3JW
England

Holland Radio
143 Greenway
Greenside, Johannesburg
Republic of South Africa

SQUIRES SANDERS, SS1R (701 series); SS1V-SS1-S, SS1-MS mint condition. Factory manuals and other parts. Heath HX-10 transmitter with accessories, HO-10 scope. Icom IC-280 FoMoXceiver, Kenwood TR-2400. All in excellent condition with manuals. (616) 382-5401. K8WPO, 9549 No. 17, Kalamazoo, MI 49007.

RUBBER STAMPS: 3 lines $4.50 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

COLLINS KWM-380 with all optional interfaces. Perfect $2495. WA6NWP, Sid, 209-642-3363. 52632 RD 426, Oakhurst, CA 93644.

KEYER/CODE trainer chip. One evening project. Great project for beginners, clubs or anyone who needs a good, low cost iambic keyer or code trainer. $15 ea. PP. VISA/Mastercard accepted. Micro Digital Technology, PO Box 1139, Mesa, AZ 85201. (602) 897-2534.

REPAIR, ALIGNMENT, calibration. Collins written estimates $25; non Collins $50. K1MAN. (207) 495-2215.

ATLAS 350XL owners group. Send QSL card with s/n your rig. Know anyone who repairs them? Have any technical information to share? Any questions? Rod, N5NM, Box 2169, Santa Fe, NM 87504.

CHASSIS and cabinet kits. SASE K3IWK.

SCHEMATICS: Radio receivers 1920/60's. Send name brand, model, SASE, Scaramella, P.O. Box 1, Woonsocket, RI 02895-0001. (602) 897-2534.

AMATEUR RADIO's newspaper — Worldradio. Latest info. One year subscription (12 issues) only $10. Worldradio, 2120-B 28th St., Sacramento, CA 95818.

TI 99/4A Random Code practice programs. Dr. Code "General" sends international Morse code and prints on screen; you choose: speed, tone, which characters to be sent, *spacing, and more! Dr. Code "Speech"* same as "General" with speech; you choose how many characters before speech check. For cassette of both copyrighted programs and conditional copying privileges, send $10.00 plus $3.00 shipping and handling to N5ESF, Rt. 1, Box 1326, Lake Charles, LA 70601.

ENJOY SATELLITE TELEVISION. Save money with easy, guaranteed, do-it-yourself antenna plans/kits. Electronic knowledge not necessary. Send $1.00 for catalog or $8.95 for "Consumer Guide to Satellite Television". GFI-41, Box 9108, Missoula, MT 59807.

VHF, UHF, ATV KITS High quality professional kits from Wood and Douglas, England. Designs for RX's, TX's, XCVR's, Pre-amps. Catalog Lance Lyman, K8IXZ, Tactical Electronics Corp., P.O. Box 1743, Melbourne, FL 32902.

ELECTRON TUBES: Receiving, transmitting, microwave . . . all types available. Large stock. Next day delivery most cases. Daily Electronics, 14126 Willow Lane, Westminster, CA 92683. (714) 894-1368.

S9 PLUS OTH. 59 acres for antenna farm — elevation above Derry repeater, equals Francestown repeater — south slope *Crotched* Mountain for wide propagation window, solar exposure and view — spacious 4-room lodge, appilanced kitchen, separate bunkhouse with shop and garage, excellent condition — paved access — privacy — one hour from Nashua or Manchester. Owner retiring, must sell. "CW" Farr, W1WMK, Broker, Greenfield, NH 03047 (603) 547-2053.

DIGITAL AUTOMATIC DISPLAYS for FT 101, and TS 520 (functional DG-5 replacement) Collins, Drake, Swan, Heath, and most others. 6½" digits. Write for information. Grand Systems, P.O. Box 2171, Blaine, WA 98230 (604) 530-4551.

WANTED: Cash paid for used speed radar equipment. Write or call: Brian R. Esterman, PO Box 8141, Northfield, Illinois 60093. (312) 251-8901.

$$$$ SUPER SAVINGS on electronics parts, components, supplies, and computer accessories. Free 40-page catalog for SASE. Get on our mailing list. BCD Electro, PO Box 830119, Richardson, TX 75083. Or call (214) 690-1102.

WANTED: Old microphones, remote mixers other misc related items. All pre-1935. Box Paquette, 107 E. National Avenue, Milwaukee, WI 53204.

R L DRAKE COLLECTION 2A receiver $100, 2B receiver & 2BQ $125, 2C receiver & 2CQ $135, TR-3 Xcvr and AC-3 $175, 2NT transmitter x/10 Novice crystals $85. All completely reconditioned, re-tubed and guaranteed. Plus UPS Ev Taylor, W7BYF, 2921 Loyola Dr., Davis, CA 95616 (916) 756-7372 eves.

RECONDITIONED TEST EQUIPMENT *$1.00 for catalog.* Walter, 2697 Nickel, San Pablo, CA 94806.

FOX-TANGO Newsletters — Since 1972, the prime source of modifications, improvements, and repair of Yaesu gear, free to Club members. Calendar year dues still only $8 U.S., $9 Canada, $12 elsewhere. Includes five year cumulative index by model numbers, or send $1 for index and sample Newsletter. Fox Tango Club, Box 15944, W. Palm Beach, FL 33416.

RTTY-EXCLUSIVELY for the Amateur Teleprinter. One year $7.00. Beginners RTTY Handbook $8.00 includes journal index. P.O. Box RY, Cardiff, CA 92007.

WANTED 220 MHZ FM TRANSCEIVER Prefer Xtal controlled W0BJ, 2801 Wright, North Platte, NE 69101.

IMRA International Mission Radio Assn. helps missioners — equipment loaned; weekday net, 14.280 MHz, 2-3 PM Eastern. Br. Frey, 1 Pryer Manor Rd., Larchmont, NY 10538.

"HAMS FOR CHRIST." Reach other Hams with a gospel tract sure to please. Clyde Stanfield, WA6HEG, 1570 N. Albright, Upland, CA 91786.

TENNATEST — Antenna noise bridge — out-performs others, accurate, costs less, satisfaction guaranteed, $41.00. Send stamp for details, W8URR, 1025 Wildwood Road, Quincy, MI 49082.

VIC-20 PHONE PATCH. Build your own simplex autopatch for less than $50 using your own transceiver and VIC-20 or Commodore 64. For full documentation and program cassette tape, send $20 to: KIE Enterprises, P.O. Box 72, Running Springs, CA 92382 (714) 867-7120.

WANTED: Early Hallicrafter "Skyriders" and "Super Skyriders" with silver panels, also "Skyrider Commercial", early transmitters such as HT-1, HT-2, HT-8, and other Hallicrafter gear, parts, accessories, manuals. Chuck Dachis, WD5EOG, The Hallicrafter Collector, 4500 Russell Drive, Austin, Texas 78745.

VERY In-ter-est-ing! Next 4 issues $2. Ham Trader "Yellow Sheets", POB356, Wheaton, IL 60189.

PRINTED CIRCUIT DRILLS: *sizes 53, 60, 68 and 70 thru 76.* $1.25 each, 10 assorted $9.95. Western Electic transformers $1.50 each. Two Way Radio, 437 Payne Drive, Endicott, NY 13760.

Coming Events

ACTIVITIES

"Places to go..."

MASSACHUSETTS: The 35th annual New England DXCC Dinner, November 10, Concord Lodge of Elks, Baker Avenue, West Concord. Starts 2 PM with a variety of DX talk and slide *programs. Admission $2.00. Cocktail hour 6 PM followed by* family-style dinner starting at 7:30 PM. Banquet $14.95. For information: Steve Tolf, K1ST, 12 Phylmor Drive, Westboro, MA 01581.

PENNSYLVANIA: The Foothills ARC's 16th annual Hamfest, St. Bruno's Church, South Greensburg, Saturday, November 3. Tickets $2.00 or 3/$5.00. Indoor flea market tables $5.00. Food, refreshments. Mobile check-in 147.78/18. For information, tickets or tables contact WA3HOL or write F.A.R.C., PO Box 236, Greensburg, PA 15601.

PENNSYLVANIA: The R.F. Hill Amateur Radio *Club's annual* indoor Winterfest, Sunday, November 4, Sellersville National Guard Armory. Doors open 8 AM. Entry $2.00. Non-ham spouse and kids admitted free. Food on premises and nearby restaurants. Vendors indoor space $6.00 each, outdoor space $4.00 each. Admits one. Bring own tables. For reservations: PO Box 29, Cofmar, PA 18915 (215) *721-0278. Talk* in on 145.19 (R), 146.88 (R) and 146.52 simplex.

MASSACHUSETTS: The Honeywell 1200 Radio Club, sponsor of 147.72/12 repeater and the Waltham Amateur Radio Association, sponsor of 146.04/64 repeater, will hold their annual Amateur Radio and electronics auction, Saturday, November 17, Honeywell Plant, 300 Concord Road, Billerica. Doors open 10 AM. Free admission and parking. Snack bar and bargain parts store. Talk in on both repeaters. For information: Doug Purdy, N1BUB, 3 Visco Road, Burlington, MA 01803.

NEW YORK: Radio Central ARC "Ham-Central" Sunday, November 25, 1984, 9 to 3 PM. Social Hall, Temple of Isaiah, 1404 Stony Brook Road, Stony Brook, NY. Seminars will be presented. For information contact Bob Yarmus, K2RGZ (516) 981-2709 or write 3 Haven Ct., Lake Grove, NY 11755.

INDIANA: The 12th Fort Wayne Hamfest sponsored by the Allen County Amateur Radio Society, Sunday, November 11, Allen County Memorial Coliseum, Coliseum Blvd. Advance tickets $3.00; $3.50 at door. Tables $8.00. Premium tables

$20.00. No table sales at door. Ticket and table deadline October 20. All classes of exams given. Send Form 610 and SASE to: V.E. Coordinator, FWRC, P.O. Box 15127, Fort Wayne, IN 46885 by October 26. Large indoor flea market and commercial vendors. The infamous Ham Band directed by Luke Matthew, WB9DWJ. Vendor setup 5 AM to 7 AM. Public 8 to 4. Talk in .88. For information, tickets, tables: Hamfest Chairman AC-ARTS, PO Box 10342, Fort Wayne, IN 46851 or call Dave Smith, KA9FFET (219) 493-2439.

MICHIGAN: The Oak Park High School Electronics Club presents a Swap & Shop, Thanksgiving Sunday, November 25, Oak Park HS, Oak Park. Donations $2.00. Tables $6.00. Refreshments. For information: SASE to Herman Gardner, Oak Park HS, 13701 Oak Park Blvd., Oak Park, MI 48237. (313) 968-2675.

OHIO: The Massillon ARC will sponsor Auctionfest 84 on November 11, Massillon K of C Hall off Rt. 21. 8 AM to 5 PM. Sellers set up 7 AM. Admission $2.50 advance, $3.50 door. Tables available $7.00/8'. Refreshments. Dinner. Auction 11 AM. Talk in on W8NP, 147.78/18. For information/registration: MARC, 920 Tremont Avenue S.W., Massillon, OH 44646. SASE please.

OPERATING EVENTS

"Things to do..."

NOVEMBER 25 AND 26: The BOMB Squad (Best of Mt. Baldy) will operate W6HCP (Hollywood Christmas Parade) from 1600Z, November 25 to 0400Z, November 26. Frequencies: 7.284, 14.284, and 21.284 MHz SSB. SASE to W6GVR for special commemorative QSL.

NOVEMBER 22, THANKSGIVING DAY: A special events station (WA1NPO) will be operating from Plimoth Plantation in the museum's 1627 Pilgrim Village from 1300 GMT to 2000+ GMT with participation of the UK Club Station GB0UST, GB2UST, GB4UST. To receive a certificate, send proof of contact and $1.00 domestic or 4 IRC's to Whitman ARC, PO Box 48, Whitman, MA 02382. For information: KA1CZS (617) 826-4772; WB1CNM (617) 586-7524. Rosemary Carroll, Plimoth Plantation, PO Box 1620, Plymouth, MA 02360. (617) 746-1622 or Peter Jackson, G3AOV, 32 Brown Avenue, Parkfield, Nantwich, Chesshire, UK Phone 0270-627149.

NOVEMBER 10 AND 11: The Armored Force Amateur Radio Nationwide Emergency Team (A FAR NET) will sponsor a Veteran's Day special event station event station from 1200 GMT Saturday to 2400 GMT Sunday. Primary frequencies: 7.285, 14.325, 21.375 and 28.640 ± QRM. For a certificate send QSL and large SASE to Alfred G. Beutler, 36 Manchester Road, East Aurora, New York 14052.

NOVEMBER 17 AND 18: VK versus the World. Sponsored by the CW Operators QRP Club. Contestants may work DX or own country for scoring. QRP stations must sign QRP for identification. 0000Z Nov. 17 to 2400Z Nov. 18. Exchange: All stations 6 digits comprise RST followed by serial number, commencing with 001 to 999 then commence again. For information SASE to Contest Manager, PO Box 109, Mt. Druitt, N.S.W. 2770 Australia.

DECEMBER 1 AND 2: The 20th annual Telephone Pioneer QSO Party starts 1900 UTC Saturday to 0500 UTC Monday. 1.8-420 MHz +. Exchange: Contact number and chapter number. ITPA Club or chapter name. Send logs showing date, time station worked, chapter name and number, contact number and claimed score prior to January 15, 1985 to: Ted Phelps, W8TP, c/o John D. Burlie Chapter No. 89, TPA, 6200 East Broad St., Columbus, OH 43213.

DECEMBER 2: "Packet Radio Overview and Prospective" will be the subject of the 2nd North American Teleconference Radio Net (TRN). Learn about packet radio from two of its leading developers by tuning into TRN at 6 PM CST (0000Z). For a complete list of gateway station locations and frequencies write TRN Manager, c/o Midway Amateur Radio Club, PO Box 1231, Kearney, NE 68847-1231 SASE please.

THE AMATEUR RADIO MOTORCYCLE CLUB NET has moved to 3.888 MHz each Thursday night at 0300Z. All brands of bikers and riders are welcome. For more info send large SASE to Gary McDuffie, Rte. 1 Box 464, Bayard, NE 69334.

THE DELAWARE-LEHIGH ARC (W3OK) will operate Dec. 21, 22, 23, 1984 on 3.990, 7.299, 14.225, 21.325 and 28.525 MHz speading Holiday best wishes from Bethlehem, PA, The Christmas City. Large SASE to colorful certificate c/o DLARC Greystone Bldg. Gracedale, Nazareth, PA 18064.

VIRGINIA FONE NET 50th ANNIVERSERY CERTIFICATE commemorating 50 years of continous operation on the 75 meter band passing traffic in Virginia is being offered by the VFN. Work 25 VFN members and send log to K4IEC with #10 SASE for certificate. Contacts must be made between 9/30/84 and 6/30/85.

DECEMBER 1984 / $2.50

ham radio magazine

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DECEMBER 1984

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contents

REFLECTIONS
the good, the bad, and the ugly

There was plenty of good news in Amateur Radio in 1984. For example, we saw the beginning of the VEC program as the FCC, per its plan, extricated itself from testing. Fears were expressed that we wouldn't be able to effectively provide our own examining program. Now, however, there are 21 VEC's doing quite a good job — particularly the Volunteer Examiners, who are essential to the success of this program. And there were those who said it couldn't be done! (I feel it's appropriate to both thank and list the VEC's currently in operation. See **table 1.**)

Band expansion: it's incredible, but it's actually occurred. There was a time when I thought the HF bands would never see the increased U.S. phone allocations, but as of September 1, 1984, the 80, 15 and 10-meter bands did indeed change — for the most part, for the better . . . although it all depends on your point of view. (25, 50, and 200 kHz respectively were added to the lower portion of the phone segments of the 80, 15, and 10-meter bands, respectively.) Some Canadian Amateurs expressed disappointment in the U.S. phone band expansion on 80 meters from 3.775 down to 3.750 MHz and indicated that it might be a good idea to completely modify *their* band plan, eliminating the CW/PHONE mode subband segments. This, of course, would affect U.S. Novice operations and could cause yet another round of band changes in the U.S. As of this writing, it hasn't happened yet, though the Canadian D.O.C. is studying the proposal. (For those who operate on 80 meters — please keep the internationally regarded DX window (3.790-3.800) clear for DX contacts only.)

Some of the news was bad. Two petitions threatening a portion of the 220 MHz Amateur band were presented to the FCC by organizations (LMCC and STI) representing the land mobile industry. This situation requires close scrutiny and action by the Amateur community. (For details, see "The Endangered Spectrum: 220 MHz Under Fire," *ham radio*, October, 1984, page 6.)

One of the decisions made at WARC '79 concerned the allocation of the upper portion of the 160-meter Amateur band or more precisely, 1.9 to 2.0 MHz. In implementing the WARC agreements, the FCC has proposed to divide the 1.9 to 2.0 MHz spectrum into two parts: reserving the lower 50 kHz exclusively for non-Amateur service and assigning the upper 50 kHz to be shared by Amateurs and commercial services. Unfortunately the frequencies of 1907-1912 are used exclusively by JA hams, and this change may eliminate any chance of working JA's on 160 meters. Nevertheless, in terms of net frequency allocations gained as a result of WARC '79, the Amateur service is *still* a winner.

There was good news for those working DX. 1984 was a bumper crop year, with a number of rare countries — for example, BV, BY, and CEOX — showing up regularly on the HF bands.

One item — an ARRL "Request for Issuance of Declaratory Ruling" submitted to the FCC — has potentially vast, beneficial implications for Amateur Radio . . . but *only if hams respond* — and respond in adequate numbers. In its request, the League asked the FCC to exercise preemptive authority over local ordinances such as restrictive zoning and emissions limitations that would inhibit Amateur operations. A weak response from the Amateur community could actually cause this request to backfire. (Remember, it's very difficult to communicate on any of the bands without an antenna *or* transmitter.) Formal comments should be filed with the FCC before December 14, 1984.

In other news, Los Angeles-area jammers tried unsuccessfully to disrupt the 2-meter radio link supporting the Olympic Torch Run. Event organizers foiled the jammers by switching to private repeaters, and the run continued without interruption.

The FCC took decisive action against malicious interference, taking one jammer all the way to Federal Court, where he was convicted and placed on three years' probation. Several other investigations resulted in license suspensions.

Legislation that would make malicious interference a criminal violation (rather than an administrative violation) is pending in both the Senate and the House. Senator Barry Goldwater, K7UGA, introduced S-2975, which would make interference with any radio service a violation of federal law; Representative Jim Bates of California introduced a similar bill in the House.

It was quite a year for those who made Amateur Radio news and for those who reported it. Special thanks are due the latter: W9JUV, W5YI, Westlink, and the many other sources — including the many club newsletters that arrive regularly — that help keep us all informed.

Rich Rosen, K2RR
Editor-in-Chief

table 1. VEC's in operation as of August, 1984.

call region	organization and location
2	Metroplex Amateur Radio Association P.O. Box 237 Leonia, New Jersey 07605
	Schenectady Amateur Radio Association, Inc. P.O. Box 6 Alplaus, New York 12008
3	Laurel Amateur Radio Club, Inc. P.O. Box 3039 Laurel, Maryland 20708
4	Mid-South VEC 2020 St. Elmo Memphis, Tennessee 28213
	Central Alabama VEC 606 Tremont Street Selma, Alabama 36701
	Western Carolina Amateur Radio Society P.O. Box 1618 Ashville, North Carolina 28816
	Charlotte VEC 227 Bennett Lane Charlotte, North Carolina 28213
	Triad Emergency Amateur Radio Club 3504 Stonehurst Place High Point, North Carolina 27260
5	Dallas Amateur Radio Club P.O. Box 173 Dallas, Texas 75211
6	Greater LA Amateur Radio Group c/o Steve L. Shafit, NE6L 21921 Lanark Street #201 Canoga Park, California 91304
	SANDARC-VEC P.O. Box 5023 La Mesa, California 92041
7	Boeing Employees Amateur Radio Society P.O. Box 3707 Seattle, Washington 98124
8	Dayton Amateur Radio Association P.O. Box 44 Dayton, Ohio 45401 (Phasing out as of January 7, 1985)
9	DeVry Amateur Radio Society 3300 North Campbell Avenue Chicago, Illinois 60618
10	PHD Amateur Radio Association P.O. Box 11 Liberty, Missouri 64068
11	Anchorage Amateur Radio Club P.O. Box 101987 Anchorage, Arkansas 99510
12	(Caribbean Insular Areas) Director, Military Affiliate Radio System P.O. Box 7388 Cidra, Puerto Rico 00639
13	(Hawaii and Pacific Insular Areas) Honolulu Amateur Radio Club 3251 Pakanu Street Honolulu, Hawaii 96822
	Koolau Amateur Radio Club 45-529 Nakuluai Street Kaneohe, Hawaii 96744
All call regions:	The W5YI Report P.O. Box 10101 Dallas, Texas 75207
	American Radio Relay League 225 Main Street Newington, Connecticut 06111

presstop de W9JUV

ACCESS TO TWO NEW AMATEUR BANDS AND CHANGES TO SEVERAL OTHERS are the subject of a WARC implementation Notice of Proposed Rule Making released by the FCC in October. The two new bands are "12 Meters" (24.890-24.990 MHz), and "33 cm" (902-928 MHz).

For 12 Meters, The FCC Proposes Dividing The Band at 24.930; RTTY would be permitted from 24.890 to 24.930, and SSB, AM, and NBFM phone plus facsimile and SSTV from 24.930 to 24.990. CW would be permitted throughout the 100 kHz-wide band. They also propose giving General and higher class licensees access to the band, with full legal power (1500 W PEP). On 12 meters, the Amateur Service would have Primary status.

For The 902-928 MHz Band, The Amateur Service Would Be Secondary to industrial, scientific and medical (ISM) operations. The Commission proposes no subbands for 33 cm. It would be available to all classes except Novice, to all emissions, with a 1500 W PEP power limit. No Amateur operation would be permitted in Colorado, Wyoming, or in U.S. possessions in Region 3. As Secondary users, Amateurs would also have to avoid interfering with Government, Automatic Vehicle Monitor, and ISM. They'd also have to avoid Region 1 or 3 users, since it's an Amateur band only in Region 2.

10.100-10.150 MHz Would Also Become A Permanent Amateur Band as a result of this NPRM. This supercedes the Special Temporary Authorization that's currently providing access to the band, putting Amateur operation Primary in the allocations tables and raising power limits to 1500 W PEP. It would be available to General and above, but limited to CW and RTTY. Effective immediately the 10.109-10.115 MHz protection slot has been deleted, but the 200 W PEP power limit will continue in effect until final action on the NPRM.

Not All The Effects Of This NPRM Are Beneficial. The Commission has further delayed implementation of the 18.068-18.168 MHz band on the grounds that U.S. Government fixed operations on these frequencies are likely to continue until 1989. The NPRM also proposes deleting the 420-430 MHz Amateur allocation along the Canadian border to protect Canadian commercial users of that band. The deleted area is a band extending 75 miles into the U.S. and Alaska from the border, and within this area Amateur operation would be permitted only with a waiver from the FCC and coordination of the Canadian Department of Communications.

Comments On The NPRM, PR Docket 84-960, Are Due at the FCC December 17; Reply Comments will be due January 16, 1985. An original and five copies are required to make a formal submission; six additional copies are required if you wish a copy to go to each of the Commissioners. Send them to The Secretary, Federal Communications Commission, Washington, D.C. 20554. As this docket covers a number of bands and issues, comments are expected to be heavy. As a result, action is unlikely until late next spring.

ALL MODES FOR 160 METER USERS HAS BEEN PROPOSED in yet another FCC NPRM released in October. Amateurs responding to ARRL's earlier petition to permit RTTY on top band had suggested adding SSTV and facsimile as well, and the Commission agreed. However, the radiolocation threat to 160's top half (November Presstop) remains a very real one.

Comments On The 160 Mode Relaxation Are Due at the Commission December 20, with Reply Comments due January 22, 1985. Refer to PR Docket 84-959.

ARRL'S PETITION TO HAVE FCC PREEMPT LOCAL REGULATION OF ANTENNAS could trigger a problem for Amateurs. Amateur response thus far to the League's request, which was designated PRB-1 by the Commission, has been sparse, but it has generated a number of comments from municipalities and municipal organizations. Their response has been, understandably, that they need more—not less—control over what goes on within their borders. As a result they'd like the FCC to not only deny the League's request, but even consider giving some formal sanction to local ordinances limiting antennas and possibly even the operation of Amateur (and other) transmitters within their jurisdictions!

Such A Turnabout Would Be A Severe Blow To Amateur Radio, and strong Amateur support for ARRL's PRB-1 is needed to head it off. Though the Comment period on PRB-1 will close before this reaches print, Reply Comments (opposing increased local control) can still be filed through December 14. Send the original and four copies to the FCC Secretary, above.

THE U.S. AMATEUR POPULATION HAS DROPPED SLIGHTLY DURING 1984, from a total of 411,479 individual current licenses in effect at the beginning of the year to 409,923 at the beginning of October. Though there was a slight peak, to just over 412,000, in the spring, the subsequent decline has been at a steady rate. Though most license classes held their own or even increased slightly during the year the most dramatic change was in the Novice ranks, which dropped more than 5000 (from 85,823 to 80,461) during the nine-month period. Though many of these were upgrades, and others early recipients of the five year Novice license who lost interest, it's disturbing more new Novice licensees have not been brought in during the period.

Confirming This Concern, An Internal FCC Analysis of 1984 license statistics shows that while 18,800 people were joining the Amateur ranks, 19,964 dropped out. In addition, while 8,829 Novices upgraded during the period, 14,484 Novices let their licenses expire!

Since The Novice Exam Procedure Had Not Changed when the Volunteer Exam Program went into effect, that should not be a reason for the decline. The FCC, understandably concerned, is considering how license structure changes might influence this trend.

comments

SSB development

Dear HR:

"The Development of Amateur SSB" by John Nagle, K4KJ, (September, 1984, page 12) did a very good job of covering the history of SSB for the present day Radio Amateur.

Two names that should go down in any historical account — Don Norgaard, ex W2KUJ, and Wes Schum, W9DYV — were not mentioned. In 1948 Don wrote a series of articles on SSB for *QST*, using Bob Dome's (W2WAM) ideas to develop a phasing type of SSB transmission and reception. Then Wes came along with his Central Electronics 10A units and kits to make the nuts and bolts of phasing SSB easily available where filters were either difficult or impossible to obtain. As one of the first two hundred hams on the air on SSB, I found them a great help.

Some of those Western Electric HF SSB transmitters John mentioned as having been put into operation in the early 1930s were still in overseas service, though a little patched up, some 40 years later — when they were sold for scrap!

Wayne W. Cooper, AG4R,
ex YN1WC, HR2WC
Miami Shores, Florida

loop antenna

Dear HR:

I think I can clear up any questions about the loop antenna (see WD6DUD letter, September, 1984, page 8) mentioned by Bill Orr in his February, 1984, column (see page 63). I've just returned from Germany with the original article that appeared in the May, 1983, issue of *CQ-DL*, the German ham magazine.

As you can see (**fig. 1**), there are actually three different methods of coupling to this antenna. C2 in Bill's diagram (**fig. 2**) is unnecessary as the feedpoint reactance can be tuned out with the loop tuning cap C1. According to diagram 13.3 of the original (**fig. 1C**), the tap should be made at 1/5 loop diameter with a spacing of 1/200 wavelength.

The length given in Bill's article is incorrectly shown in the diagram (**fig. 2**). The correct length is 0.2 wavelength at the *highest* frequency. The lengths given in the table (February, page 65) are OK.

The author states that the SWR is below 1.5 to 1 over an octave frequency range, but this will require motor tuning of the loop tuning cap because the bandwidth at the low frequency end of the range is very narrow (6 kHz on 80 meters). As for the groundplane dimensions, he recommends that it extend out at least twice the loop diameter, but further if possible. As with any small loop, the radiation resistance is quite low and ground losses can eat up a lot of signal if not minimized. This is also true of conductor losses, making the use of copper tubing a necessity. I would recommend 1/2-inch tubing because this size is readily available at plumbing and heating stores and will have less loss than the 3/8 size mentioned. Don't try to save money by eliminating the ground return line for the same reason.

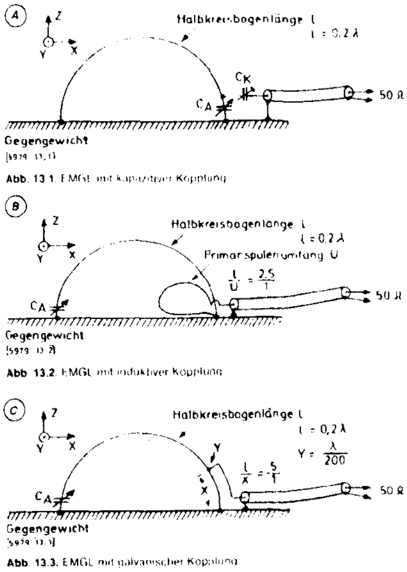


fig. 1. *CQ-DL* illustrations show three different methods of coupling.

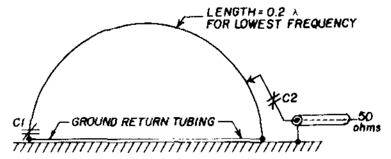


fig. 2. DL2FA groundplane loop as shown in W6SAI's February column.

Roger D. Johnson, AD1G
Caribou, Maine

spectrum history

In the interest of brevity, several lines of Joe Schroeder's (W9JUV) segment of the October editorial ("The Endangered Spectrum: 220 MHz Under Fire," page 6) were omitted. They are reproduced here in the interest of historical accuracy — **Editor**.

When the U.S. entered WW II the 1-1/4-meter band was just that: 224-230 MHz, a perfect harmonic of 2-1/2 meters (112-116 MHz), 5 meters (56-60 MHz) and so on down. A few years earlier, however, the entire spectrum above 110 MHz had been designated simply "Amateur and Experimental."

When the war ended and Amateur operations resumed, the 1-1/4 meter band was temporarily shifted to 235-240 MHz (probably to avoid some of the early radars still active between 180 and 220 MHz). Then, in the late 1940s, it was finally shifted to the present 220-225 MHz slot.

short circuit

Guerri Report

In "The Guerri Report," (November, 1984, page 54), the line that reads, "A complete SSB receiver could actually consist of 3000 individual receivers, each having 1 kHz bandwidth . . ." should be corrected to read "1 Hz." This suggests the possibility of 3000 discrete 1-Hz spectrum segments comprising a single voice bandwidth channel, similar to the technique utilized in spread spectrum applications.

ham radio

complete solar power
for your ham station

Clever parts selection and careful siting deliver economical alternative power

We've all noticed, with increasing alarm, the steep rise in our monthly electric bills. Here in Hawaii, the rates have passed 13 cents per kilowatt hour, and show signs of continuing upward. Being of a prudent nature, I decided to make the effort to reduce consumption by all practical means. Switching to roof-mounted solar collectors for hot water heating was first. Converting my station to solar power was next.

Photovoltaics (PV) offer an attractive solution to the high cost of electric power if one can afford the initial capital investment. Friends in this industry claim that suitable panels shouldn't cost more than $8 to $10 per watt; this price, however, is based on a quantity order. I figured that a single panel of approximately 30 watts would be adequate for my purposes. The only problem would be to find a vendor who would provide a suitable panel for approximately $300.*

After contacting all the sources I knew, I found that such a unit might normally retail for as much as $500, more than I wanted to spend. Luckily, a California manufacturer agreed to provide one unit for $350. This particular panel (SOLEC No. 4136) provides about 2.5 amperes at 18 volts in full noon sun.

*For a list of manufacturers, send an SASE to *ham radio*, Greenville, N. H. 03048.

The next problem was to find a suitable 12-volt battery. I wanted one that would be able to stand the heavy transmitter load without balking, and could do so for a long time. A Sears Roebuck Deep Cycle Marine Battery ($76.00) was selected.

Because the PV panel generates more voltage than the battery needs for charging, a regulator is necessary. While the panel puts out 18 volts in full sun, as noted previously, it generates less power as the sun moves lower in the sky, thus making a "lossless" regulator desirable. (Series regulators usually lose a volt or more; this loss cannot be tolerated in this application.) An excellent on/off type regulator described by Millard[1] was selected. This circuit senses when the battery is fully charged and switches off the PV input. When the battery level drops, the regulator switches the input from the PV back on.

As depicted in **fig. 1**, extensive metering is used throughout. The PV panel, regulator, and battery output are protected by circuit breakers. While low voltage DC breakers are not easy to find, a variety of surplus units are available at a reasonable price (see parts list, **fig. 1**).

A series diode is used to isolate various elements of the total unit. We are all aware of the usual approximate 0.7 volt drop across the average silicon diode. As noted previously, we are scrambling for even fractions of a volt if the PV panel output is to be used effectively. Conventional diodes — just as conventional series regulators — have excessive losses, and therefore should not be used. Power Schottky diodes with their approximate 0.2 volt drop are preferred (see **fig. 2**).

Because a charging battery generates hydrogen gas, which can easily explode, the battery should be installed outside the house in a location in which air circulation is unimpaired. Of course, the further it is from

by William L. Schreiber, NH6N, 73-4327 Imo Street, Kailua-Kona, Hawaii 96740

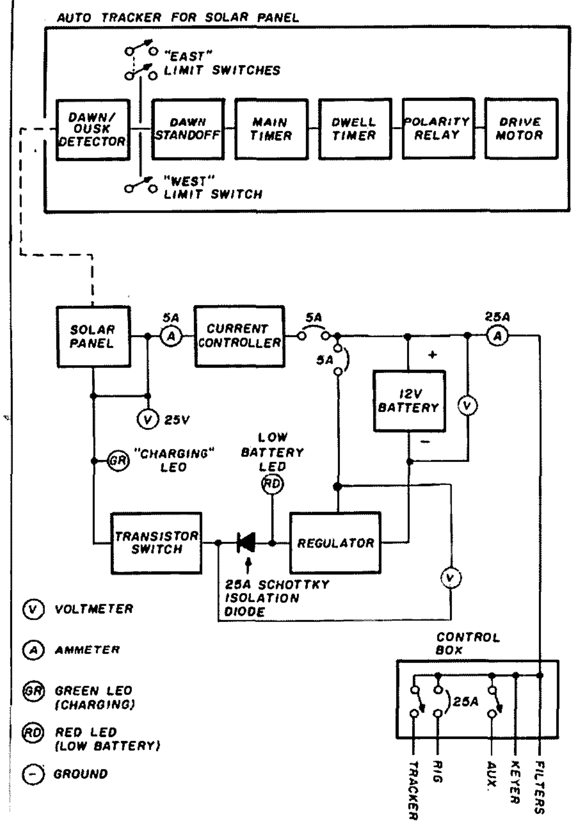
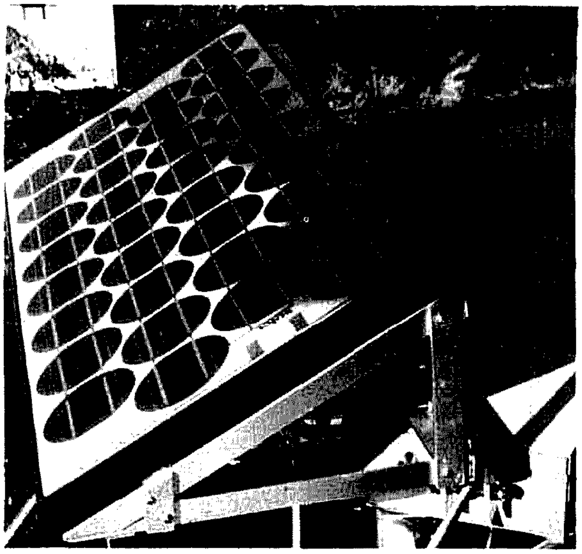


fig. 1. Block diagram of the solar power unit.

Parts suppliers

solar panel:	***SOLEC International***
	12533 Chadron Avenue
2.5 amp/18 volt	***Hawthorne, California 90250***
battery:	
12 volt deep cycle Marine battery Catalog No. 28H9652N	***Sears Roebuck***
breakers:	
surplus low-voltage DC breakers	***Fair Radio Sales Lima, Ohio 45802***
Schottky diode: SK 9151	***Quement Electronics 1000 Bascomb Avenue San Jose, California 95128***
bearings	***Edmund Scientific 101 East Gloucester Pike Barrington, New Jersey 08007***

the rig and the PV panel, the more critical the choice of cable becomes. I chose No. 10 copper wire which also proved to be quite expensive — because of its



Top front view of PV panel and drive control box.

tolerable IR loss. On a per-foot basis, a set of automobile jumper cables (copper) costs less than a third the price of regular wire. So I decided to use these between the battery, the control box, and the rig instead. But my PV panel is about 25 feet (7.62 m) from the battery — much longer than any jumper cables I could find. So I found, in my parts stock, a length of coaxial cable, similar to RG-8/U, that proved to have a No. 10 inner conductor. I used this for the run from the battery to the panel.

mounting the panel

Mounting the PV panel proved to be a problem. It was obviously important to be able to rotate it and even change its elevation if necessary. After perusing EPA documents on solar hot water panels, I decided that a good compromise for the elevation angle was to make it about 10 degrees plus local latitude. My location in the southern part of the Hawaiian Islands is about 18 degrees North latitude. So a 28 degree fixed elevation was selected. I built a frame out of 1 inch (2.54 cm) aluminum angle stock, and bolted it together with stainless steel hardware.

My first attempt to rotate the assembly involved mounting it on a wide-flange bicycle wheel hub. I bent a 1 × 1/4 inch (2.54 × 0.64 cm) sheet of flat aluminum stock to resemble a bicycle fork, and the PV frame was bolted to it. The thing worked, but it was an awkward arrangement and didn't seem sturdy enough to survive our occasional tropical storms.

I finally found "lazy Susan" bearing assemblies in various sizes. A 12 inch (30.5 cm) and a 6 inch (15.2 cm) bearing were secured. The larger one was bolted to two 12 × 12 inch (30.5 × 30.5 cm) 16-gauge aluminum sheets, with the PV frame bolted to the top

sheet. A low-friction bearing resulted, and the whole assembly could be rotated with light pressure (fig. 3).

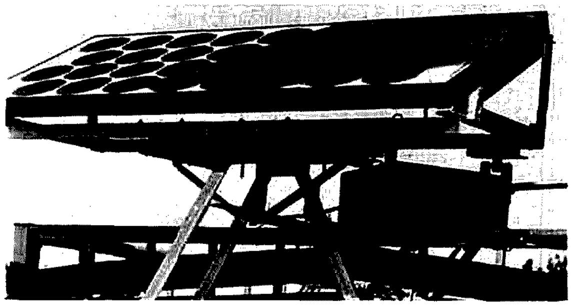
Total rotation angle is an additional consideration. Because very little current was developed by the PV panel for the first hour or so after sunrise, I decided to limit rotation between Southeast and West for approximately 135 degrees of rotation.

With these successes in hand, it became obvious that a motor-driven tracker should be added so that the panel would always face the sun. Two basic choices were considered. The first was a true tracker based on two phototransistors. It was breadboarded and found to work, but it had two basic flaws. A ±12 volt supply at 2 amperes was required. The battery would supply the +12 volts quite easily, but the negative voltage was not so easily provided. Since my goal was to be able to depend totally on the battery, no outside voltage could be considered. It's possible to generate −12 volts from the +12 volt battery via a power op amp inverter, but this seemed to be doing the job the hard way. In addition, an auto tracker gets into trouble when the sun goes behind the clouds for an hour or so. In order for the panel to find and face the sun, an "acquisition" system must be included in the tracker design. This was far too complicated and was summarily rejected.

The second choice was a timer-controlled system (see fig. 4). (Some circuits in Walt Jung's *IC Timer Cookbook*[2] are suitable.) My approach was to use an astable to handle the longer times — 40 to 60 minutes — and use its high output to trigger a monostable, which would have an adjustable duration of 0 to 30 seconds. This would turn the motor on via a relay.

For convenience, the timing resistors for the astable — eight 10-megohm 1/4 watt — were wired across a DIP switch. This allowed up to eight resistors to be placed in series. The timing capacitor was composed of five each 2.2 μF mylar capacitors in parallel.

At this point, it is important to consider the various pecularities of IC timers. My old standby, the 555/56, has serious limitations in applications in which time



PV panel in test position.

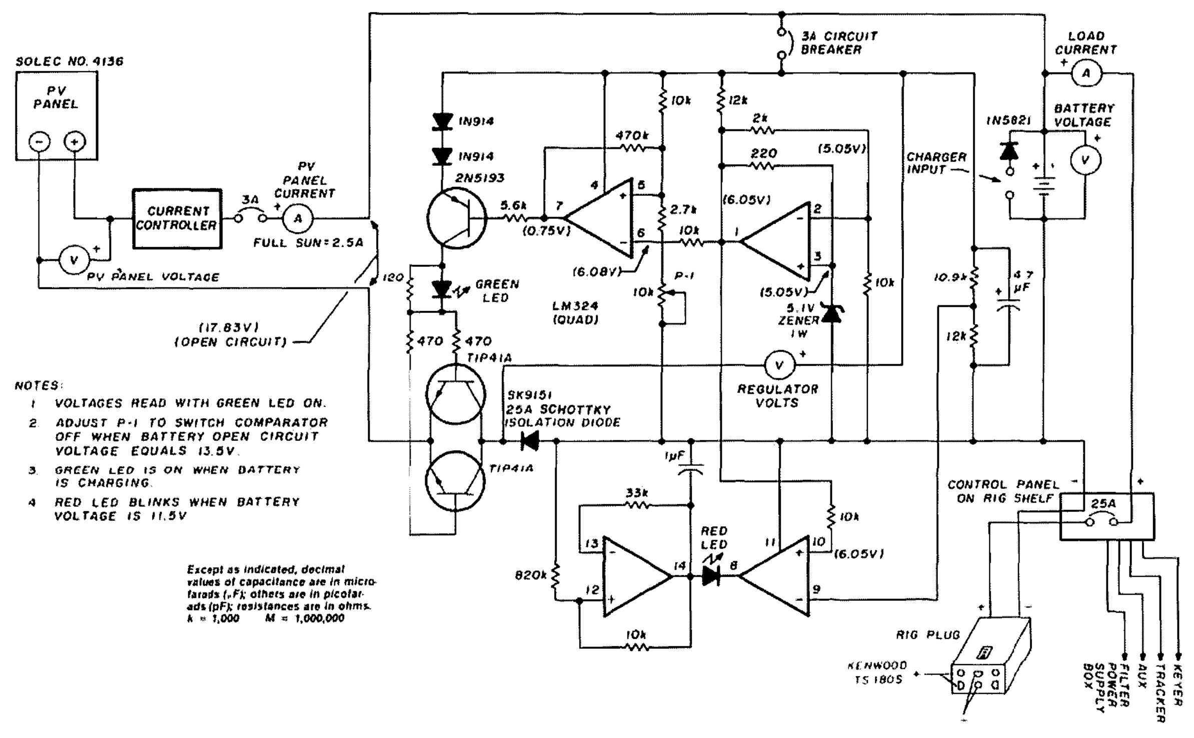


fig. 2. Solar panel and battery supply electronics.

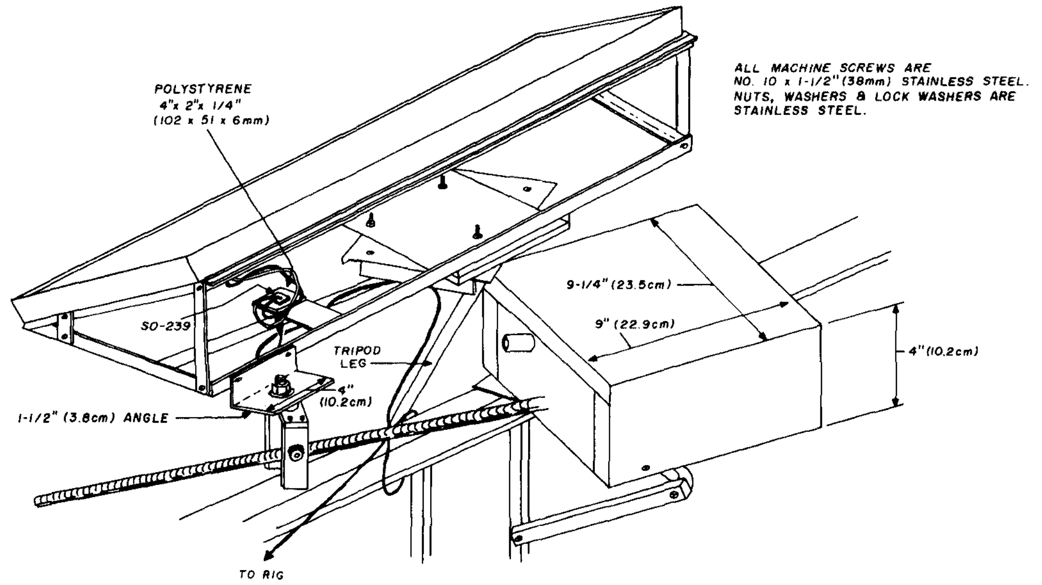


fig. 3. Low friction bearings ease rotation of the solar panel.

delays of long duration are desired. Also, the capacitors have to be carefully chosen to avoid leakage. Aluminum electrolytics are unsatisfactory, and even tantalums have to be used with caution. The low voltage film types (mylar, polystyrene, and polypropylene) are the best choice. If necessary, a quality tantalum may be used.

My first attempt at a 30-minute astable with a 555 proved to be very unpredictable. One of the problems was leakage within the IC. The other had to do with the inability of the 555 to time out with a resistor much larger than 15 megohms. Jung points this out in the *Cookbook* and suggests using the CMOS version of the 555/56 — the 755/56 made by Intel. This very low-leakage device easily handled a 100-megohm network, and was chosen for the astable.

When this was done, I noticed that after about 40 megohms were switched in, the circuit "timed out" the same even up to 100 megohms. The timing was erratic also. I learned a lesson here: the perf board was liberally splattered with rosin and fine solder splashes that are normally no problem. But throw in a lot of megohms, and the leakage both here and across the switch contacts will interfere with the timing. The solution was to clean the board with a flux solvent and get rid of the DIP switch. The bank of resistors now hangs in the air directly from the IC socket pins.

The monostable had no such problems because of the short timer delays involved. In the interest of simplicity, a dual IC — 7556 — was used for both purposes.

A practical problem incident to the 7556 is its low current sourcing capability — approximately 50 mA. I was unable to buy a multipole relay with coil currents below 80 mA here in Hawaii, so I added the Darlington power transistor as a relay driver.

dawn/dusk circuit

In the overall system concept, it was obvious that the tracker should stop when it finished its rotation toward the west. A limit switch did the job when an adjustable stop came in contact with the switch arm. Suitably connected, this turned off the motor. A means had to be provided to recognize this fact as evening approached, and the switching system allowed the PV panel to rotate back, face East, and then stop. The stopping was accomplished by two additional limit switches and an adjustable stop. At dawn, the limit switches had to be bypassed so that the timers could start the intermittent rotation.

The circuit chosen was based on the large change in resistance by a cadmium sulfide cell. Mine has a "dark" resistance of about 100K. Its "light" resistance was about 50 ohms. This characteristic was used in a simple comparator circuit. Again, the 741 op amp did not handle the relay coil current, and a Darlington power transistor was used as a booster. Should you find that the circuit "hunts" as it gets dark, it may be necessary to add hysteresis so that the op amp will have a "dead zone." This can be done by providing a very small amount of positive feedback by connecting a 6.8 Megohm 1/4-watt resistor on the comparator

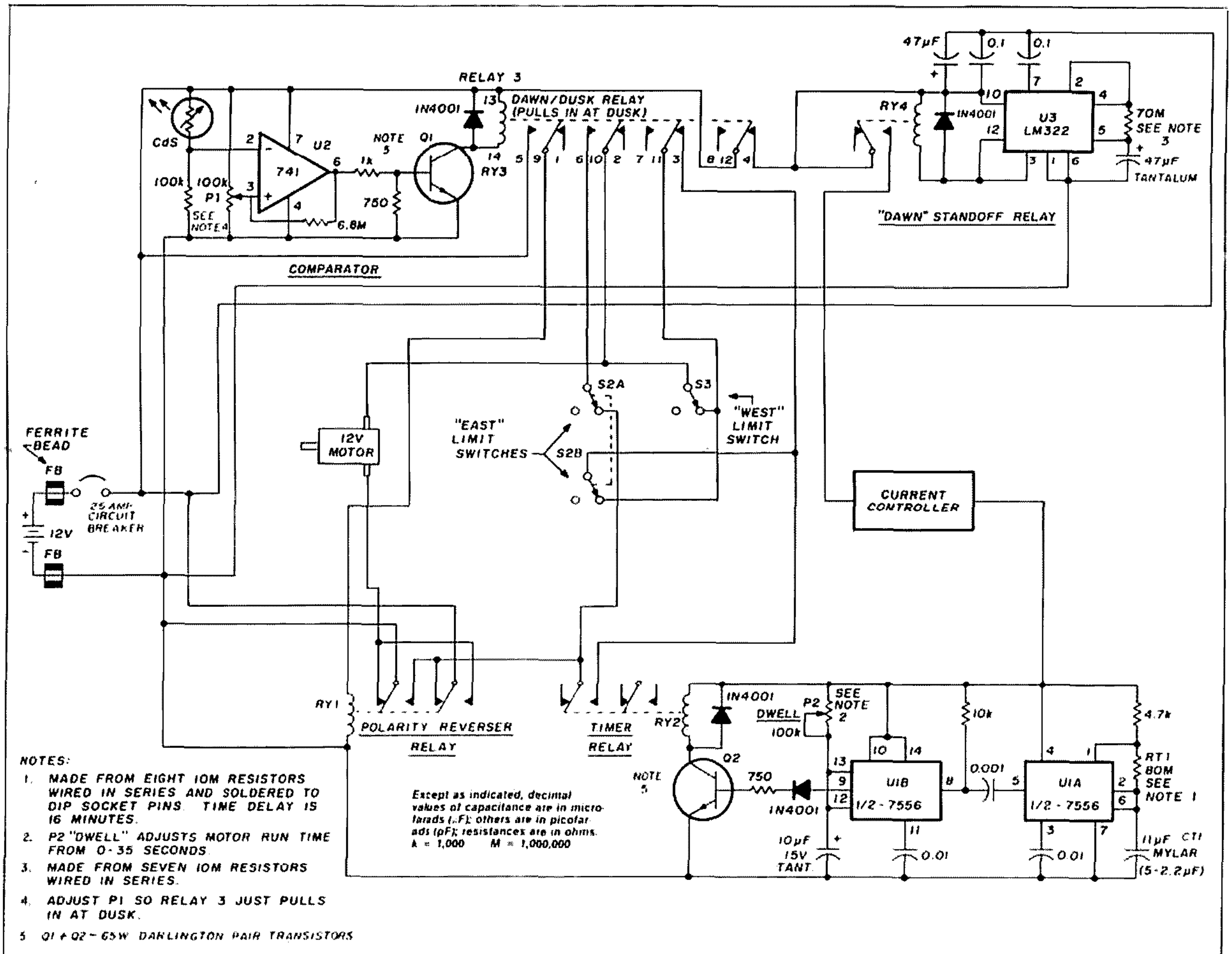
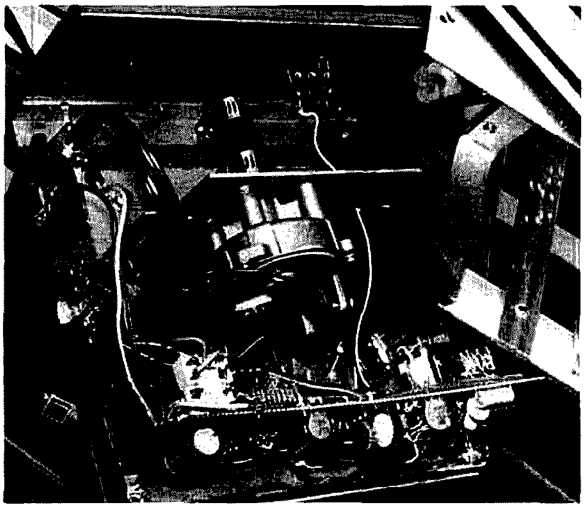


fig. 4. Autotracker circuits account for rotational limits, panel position reset and "dawn standoff" compensation for a low output condition at sunrise.



Internal view of drive control electronics box illustrates location of CdS cell (upper right), motor mount, flexible coupling and west limit switch below the wiper motor.

op amp output to its non-inverting input (pin 1 to pin 3). Another factor to consider is to match the dark resistance of the CdS cell with its series resistor. In my case, it's 100K.

dawn standoff

A number of tests revealed that the PV panel had very little output early in the morning, and reached its average output about an hour after sunrise. Because the "dawn" detector triggers at sunrise, the PV panel tends to begin rotating too early for optimum sun tracking. Adding the "dawn standoff" timer handled the situation nicely. Briefly, it uses a National LM-322 CMOS IC as a long-duration timer in a one-shot, power-up circuit.[3] The particular R_t C_t (80 Megohms, 47 microfarads) used here gives a delay of about 65 minutes. Triggered once by the "dawn" circuit, it times out, goes high, and actuates the relay that supplies power to the main timer. The delay can be adjusted over a wide range by increasing or decreasing

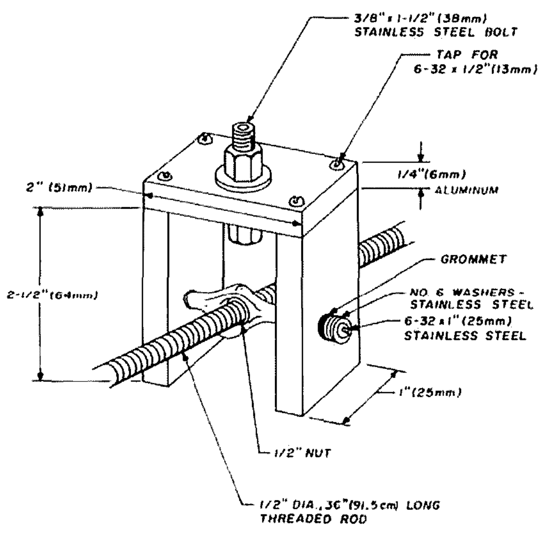
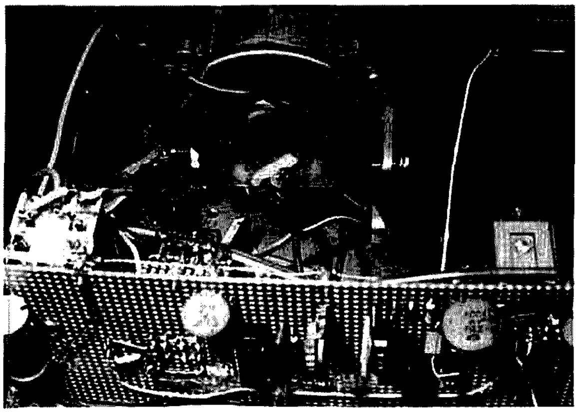


fig. 5. Mechanical details of the lead screw system.



Close-in view of drive control box electronics shows east limit switches below gear head.

C_t and/or R_t. As noted previously, the series resistor bank hangs in the air, and is soldered directly to the IC socket pins.

mechanical considerations

As can be seen in photos A and B, the electronics/drive motor box is offset to the rear on the "east" side of the PV panel. The exact location of this unit has to be determined by trial and error, and is based on the desired total rotation angle, as well as mechanical support practicalities. The center of my box is 13 inches (33 cm) from the center of the PV panel and is offset 7-1/2 inches (19 cm) to the side.

My choice of a lead screw system was limited to an available 3 foot × 1/2 inch (91.5 × 1.27 cm) (NC) steel drive rod (see **fig. 5**) which was the only threaded rod I could find in my local hardware store. An aluminum rod of the same dimension would have been preferable.

The only outside mechanical work needed was to weld a pair of threaded steel spacers (8/32 × 1 inch) (6.4 × 25.4 cm) to a steel 1/2 inch (1.27 cm) (NC) nut. A U-shaped bracket was made of 1 × 1/4 inch (25.4 × 6.4 mm) flat aluminum stock into which the nut and trunnions were fitted. The bracket was drilled at the top for a 3/8 × 1-1/2 inch (9.5 × 38 mm) stainless steel screw and nut that permitted it to be attached to an aluminum angle bracket. This in turn was attached to an extension in the PV panel. The whole arrangement resulted in a universal joint tolerant of my sloppy mechanical alignment.

Rotation angle to the west is limited by the threaded rod's contact with the PV panel structure. To the east, the system will not work beyond the point at which the rod comes into line with the center of the panel. This allows about 135 degrees of rotation. If more is required, a longer rod is needed, and the box would have to be moved out a greater distance from the PV panel.

I needed a reasonably powerful, geared down, 12 VDC motor that was both cheap and very small. Knowing that cars use all sorts of small electric motors, I went to the local wrecker and bought a nearly new 2-speed Mazda windshield wiper motor for about $3.00. It checked out at a slow speed of approximately 120 RPM — too fast for my purposes — but a lead screw system with a speed of much less than 1 RPM slowed the panel rotation down. I'll admit that it's a clumsy looking thing, but it works well and is very rugged. If you can find a small, powerful 1 RPM motor, you can probably drive the PV panel directly with a short vertical shaft and thus simplify things considerably.

motor

Since it's not possible for me to anticipate which particular motor you may use, there's no way I can indicate specific connections. In any case, the leads should be paired out with an ohmmeter. I used my bench power supply to try out my choices. A lower than normal voltage was used, and the built-in current limiter provided added protection. In any event, proceed cautiously, and determine which two leads coming from the motor (mine had six) are the proper ones.

electronics/drive box

This is made of double-sided circuit board soldered together for the top and sides. The front and bottom are 16 gauge aluminum. (The latter could have been made of the circuit board material just as well.) A small

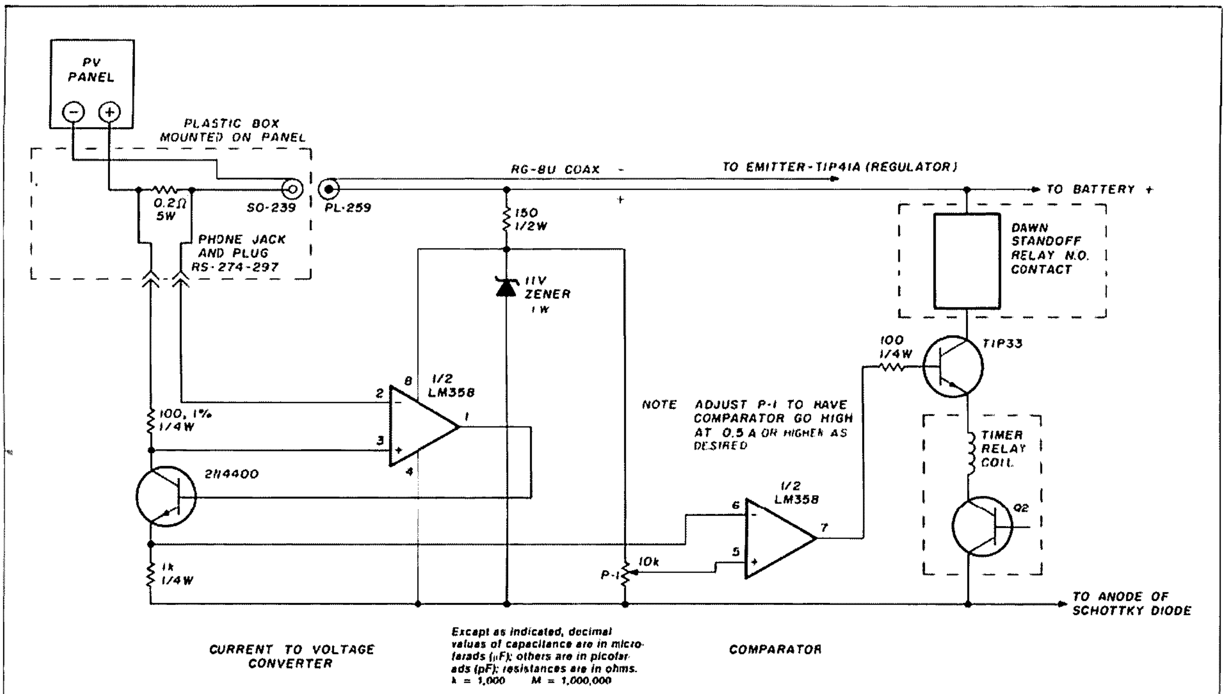


fig. 6. Current controller recognizes when a suitable charging level is achieved and stops panel rotation until the level drops.

"lazy Susan" bearing attached to the bottom panel allows rotation of the box as the panel swings.

The windshield wiper motor is mounted on an aluminum angle bracket and attached to the threaded rod with a piece of rubber hydraulic tubing that fits the motor shaft and rod tightly. The rubber tube is held in place with two hose clamps. This results in a flexible coupling which, together with the universally mounted nut and the rotating box, is very forgiving of inaccurate mechanical alignments (see photo).

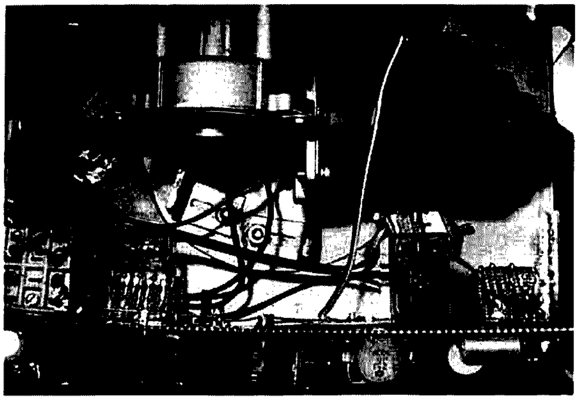
A perf board is used to mount the relays, ICs, and assorted components for the timing circuits. The lead sulfide cell is mounted in a rubber grommet on the front of the box, where it has a good "view" of the ambient light.

mounting

Mount the PV panel where it has an unimpeded "view" of the sky between southeast and west. Mine is mounted on a 10 foot × 3 inch (305 × 7.6 cm) plastic sewer pipe. Its base is set in concreted rocks, and the top has a flange that supports the 3/4 inch (1.9 cm) plywood base. Because the pipe is somewhat limber, I used guy wires to stabilize its position.

adjustments

A number of adjustments are necessary before the system is fully functional. Mechanical settings of the box position and rotation angle have already been mentioned. Electronics items needing adjustment include the following:



Edge view of control box details limit switch stops, resistor banks ("in the air") and drive motor.

Voltage regulator. The comparator pot should be set to have the green LED turn off at a battery voltage of 13.5 volts. The ideal way to do this is with a metered power supply substituted for the battery wherein the supply is set to 13.5 volts and the pot turned until the LED goes out. Alternatively, the battery can be used with an accurate voltmeter. Once the battery reaches the desired voltage, this adjustment can be made. Of course, you'll have to wait for the battery to charge up to this voltage, which may take some time.

Main timer (first half of the 7556). Depending on how long a time delay is desired, both the timing capacitors and resistor bank can be added to or subtracted from. More resistance and/or capacitance increases the delay.

Dwell (second half of the 7556). The pot allows an adjustment of 0 to 30 seconds. In my case, the main timer times out at about 12 minutes, and the dwell is set for about 6 seconds. This combination is reasonably close to keeping the PV panel pointed at the sun (± 15 degrees).

Dawn/dusk relay. The comparator's pot permits a wide range of light levels to be selected. Mine is set so that "dusk" is recognized just at sunset.

Dawn standoff. Again the RC combination controls this timer. Mine times out after approximately 1 hour after sunrise, which seems to be the point at which the PV panel produces a worthwhile output. If you want a shorter period, then decrease the resistor or the capacitor value.

Limit switches. Because it wasn't possible to find a double-pole double-throw micro switch for the "east" limit, two SPDT units were bolted together with 2-56 screws 1-1/2 inches (3.8 cm) long (photo D). The "west" limit required only one SPDT switch. Location of the stops for these switches took some careful planning to avoid mechanical interference (photo E). The whole assembly was rotated to the desired positions, and the bottom plate marked under the desired switch arm. It was then drilled and tapped for a 10-24 stainless screw 2 inches (5 cm) long. A jam nut was threaded on and tightened after the screw was adjusted up or down to make proper contact with the switch arm.

RFI

My whole assembly is quite close to the dipoles used in my ham set-up. RFI from the transmitter to the timer ICs could have been a problem. To reduce this possibility, ferrite beads are used on all wires going in and out of the control box; bypass capacitors are installed, and the whole assembly is grounded. Happily, all this worked and there is no interference problem.

a current-to-voltage converter

If you want to add some "smarts" to the system, you may want to provide a means of stopping the timing when the charging level is sufficient. I chose a level in excess of approximately 0.5 amp as the cutoff point.

This scheme was implemented by adding a series resistor of 0.2 ohm in the positive line coming from the PV panel (see **fig. 6**). One-half of a dual IC (LM-358) was arranged so that the non-inverting input received the full output of the PV panel. With a small signal transistor in an emitter follower connection, the combination gives a voltage output which is proportional to the drop across the 0.2 ohm resistor.

The other half of the LM-358 is used as a comparator. Its reference is the zener regulated panel voltage, and is adjusted by the 10K pot. It is set to change state when approximately 0.5 amp is provided by the panel. If the level is lower than this, the comparator goes high and switches on the TIP33 gate transistor, which allows the supply voltage to go to the main timer. Rotation of the panel then continues. If the level is in excess of 0.5 amp, it stops timing (and rotation) so that the system will work efficiently.

Should you want a higher or lower charging level, it is simple to arrange. Just reset the comparator's pot.

references

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2. Walter Jung, *IC Timer Cookbook*, 2nd Edition, Howard W. Sams, Inc., 1981.
3. *Application Note AN97*, National Semiconductor Corp., 2900 Semiconductor Drive, Santa Clara, California 95051.

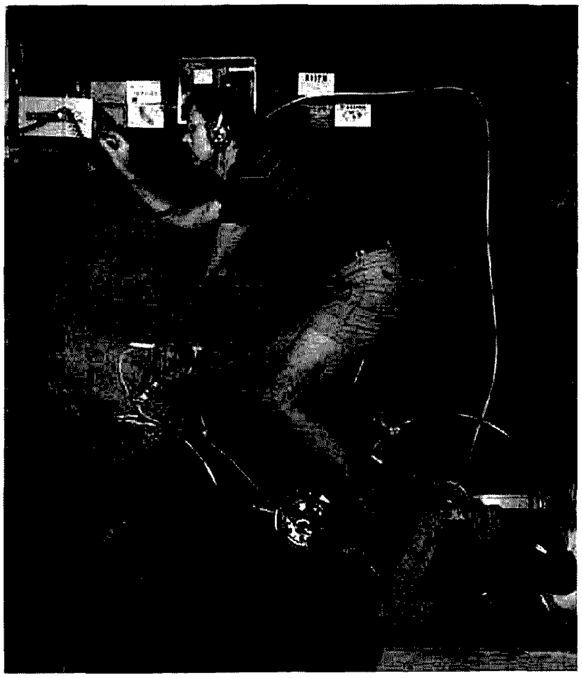
ham radio

the bicycle-powered station

All you need is some simple circuitry, a junkyard alternator, and a ten-speed bike

By combining human energy with a ten-speed bike, a used automotive alternator, and some simple circuitry, Amateurs interested in exploring alternative sources of power can put a station on the air with electricity derived from a low-cost system of only modest complexity.

The automotive alternator — readily available, inexpensive, and capable of producing up to half a kilowatt of power — is the perfect starting point for experimenters interested in producing their own electricity. This bicycle-driven power plant, designed around the common Delco-Remy alternator, provides 110 volts of regulated AC at levels up to 150 watts. Designed primarily to run a 100-watt transceiver, the setup is used several times a week as an entertaining form of physical exercise. For that reason I intentionally omitted any means of energy storage, either electrical or mechanical. (The alternator and its controlling electronics can easily be adapted to other sources of mechanical power if desired, so the circuitry should also be of interest to Amateurs interested in inexpensive wind-, water-, or gasoline-powered installations.)



The bicycle powered 100-watt Amateur station. The bicycle's rear wheel is chain-coupled to the automotive alternator, which in turn powers the transceiver. There's no electrical energy storage and no battery backup. Mounted over the operator's head are the control electronics and step-up transformer which produce the system's regulated 110-volt AC output. (*Photo courtesy W1GSL.*)

As a power source, the bicycle presents some interesting design challenges. Because the operator provides 100 percent of the station power, high efficiency is a must if over-the-air huffing and puffing are to be avoided. There is also a large variation in the amount of power delivered during each pedal revolution, so good regulator action is necessary to hold the output constant. These factors combine to make the *bicycle a difficult source of mechanical power to* apply successfully.

The performance data in **fig. 1** through **4** shows how well these problems have been solved. Mechanically, the system is somewhat noisy but still quite smooth. Its overall efficiency is such that a middle-aged adult in average condition should be able to produce 50 watts continuously for an hour without undue strain.* (A typical 30- to 40-minute radio contact leaves this operator damp but by no means exhausted.) The electronic regulator smooths out pedal-rate variations so they result in only a 2 to 3 percent modulation of the system output. In terms of transient response and spike suppression, the system is probably better behaved than the more expensive gasoline-powered generators so often used by Amateurs.

the alternator as a power source

Before discussing the design of this system, it's worth considering the alternator itself and several possible ways it can be used. The circuit diagram of the Delco-Remy unit I used (before and after modification) is shown in **figs. 5A** and **5B**. Basically, my alter-

*Consult your physician before undertaking any form of exercise program — Ed.

By **Penn Clower, W1BG**, 459 Lowell Street, Andover, Massachusetts 01810

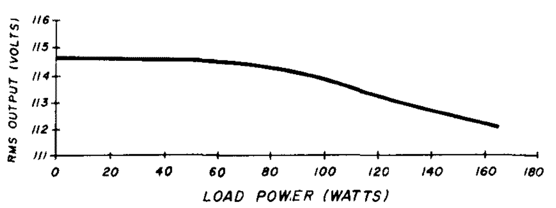


fig. 1A. Output RMS voltage regulation as a function of resistive load power. The total voltage drop is only 2 percent for a 160 watt increase in output power.

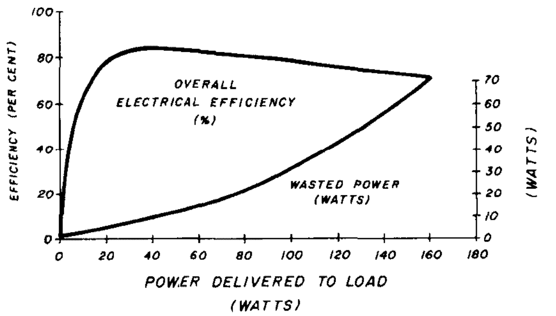


fig. 1B. Electrical efficiency. The power not dissipated in the load is lost as heat in the alternator stator and transformer resistances or is used to excite the field winding.

nator is a three-phase machine with a Y-wound stator driven by a rotating field coil. In normal use the three output phases are full-wave rectified to produce DC output with approximately 14 percent ripple. This output level is adjusted by controlling the field excitation current with an external regulator.

The windings are arranged so that seven electrical output cycles occur during one mechanical shaft rotation. This accounts for the relative advantage an alternator has over the common two-pole DC generator in low-speed applications. For a given field current, the unregulated output voltage of either machine is directly proportional to shaft rate, but the 7-to-1 electrical advantage of the alternator means it can reach the desired output level at a much lower mechanical speed.

In automotive use the regulator design can be quite crude because the car's storage battery does a good job of filtering the alternator's output. A more sophisticated regulator is necessary when the unit is used as the primary power source for a system having a minimum of energy storage capability. Overall regulation should limit the output level to within 5 to 10 percent of nominal, and care must be taken to limit the amplitude of any output spikes.

Lack of energy storage will also create another problem not found in the automotive environment. During normal operation the regulator circuitry will be powered from the alternator output. At startup there isn't any appreciable output, so a low-level source of regulator power must be provided along with a means of smoothly shifting back to internal power once the system is stabilized.

Typical car alternators can produce output powers on the order of 400 watts. It's worth noting that the internal resistance of the unit used here is 0.286 ohm, so at that output level the alternator I^2R losses are almost 300 watts. This explains why the unit is normally force-ventilated by a fan. Some care must be taken, particularly if the alternator is to be bicycle driven, to minimize that wasted power. Since the internal resistance is a given constant, the only way to increase efficiency is by lowering output current, and that means either lowering output power or increasing output voltage.

application techniques

With that as background, let's consider two possible ways of using the alternator. Since much of the newer radio equipment is capable of operating directly from 12 volts DC, one immediate suggestion is to simply use the alternator as a source of low-voltage DC power. Another approach, the one used here, is to tap into the AC portion of the alternator circuit with a transformer and bring the output level up to 110 volts. Each of these choices has its merits and disadvantages.

While I didn't explore the DC approach in depth, one problem that might be anticipated with that method is worth mentioning. When used without some sort of energy storage, the DC output of the alternator will have a ripple component equal to 14 percent of the peak level. Assume the decision is made to filter that ripple down to 0.5 volt peak-to-peak (3.8 percent) when the alternator is driving a 100-watt load. At the expected ripple frequency, an 8000μF capacitor will be required to do this, and the addition of that capacitor can complicate the control loop. The capacitor gets charged through the low output impedance of the alternator, but it must get discharged through the unpredictable and often variable load resistance. This means that the regulating control loop will respond differently to overvoltage and undervoltage transients. It might be difficult to design a circuit that would work well in all situations, and certainly some extra attention would have to be paid to the issue of loop stability.

The AC approach avoids this problem and has, in addition, the advantage of being able to run just about every piece of equipment in the station. The AC frequency will vary and is typically near 300 Hz, but that's

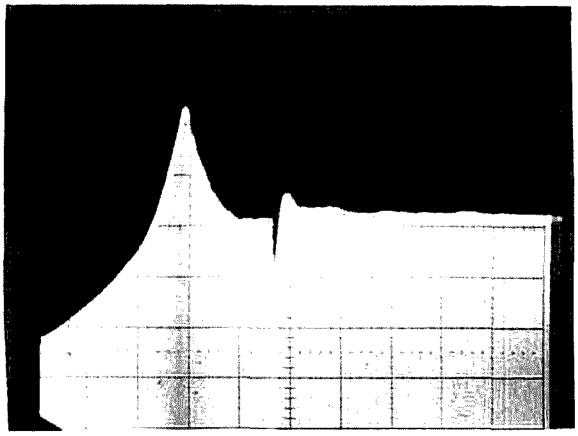


fig. 2A. The system startup transient. The AC output waveform is shown at 50 volts and 0.2 second per division; the zero level is 3 lines below the 'scope centerline. For this test the unloaded system was brought up to speed as rapidly as possible. Note the large overshoot which builds up before the regulator circuit can gather its wits. The small negative going spike at midscreen occurred when the battery driven startup current was removed.

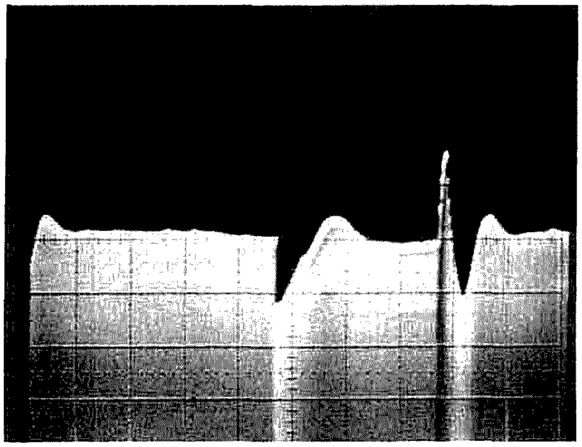


fig. 2B. Output voltage change caused by the sudden application (at midscreen) and removal (3½ divisions later) of a 75-watt resistive load. Scaling is 50 volts and 0.1 second per division, and the zero volt level is 3 lines below the centerline. This is the combined reaction of the regulator and bicycle operator; the regulator itself is even better behaved than the picture would indicate. The positive spike which occurs when the load is removed has been considerably shortened by the action of the neon bulbs.

of little consequence unless the equipment powered includes a clock or fan designed for 60 Hz. For most applications a valid argument can be made that higher input frequency will result in lower transformer core losses and improved power supply filtering. Furthermore the use of a step-up transformer allows the designer to select the alternator output voltage independently of the system's 110-volt output level. This means it's possible to set the system's output impedance for maximum efficiency at the expected power levels.

The biggest disadvantage of the AC technique has to do with the system output impedance and the nature of the load currents. Power supplies do not look like resistive loads because current is drawn only from the peaks of the rectified sine wave. Not only does this mean that current flows in pulses, but because of duty-cycle constraints the amplitude of those pulses is from four to six times the level of the DC current leaving the supply.

A small system like this is going to have a relatively large output impedance, about 24 ohms, and currents flowing through that impedance will cause a drop in the output voltage. Of course the regulator's job is to compensate for output voltage drops by increasing the field current, but unfortunately the regulator's bandwidth is only about 15 Hz. While it's perfectly capable of removing most of the effects of something like a resistive load being switched on and off at a few Hz, the regulator is powerless to do anything about the large, twice-per-cycle current pulses drawn by a transmitter's power supply. The frequency of that disturbance is simply too high. In such a case the controller may very well regulate the output RMS level as intended, but the wave form of that output will degenerate gradually from a sine to a square wave as the pulsed load currents increase.

That distortion might or might not have an effect on the radio used as a load. My transceiver is an older one with tube-type driver and amplifier stages. The filaments see no change because they react only to the constant RMS value of the supply voltage. The B+ does drop a little during a transmission though, and the peak input power to the transceiver is somewhat lower than when operated from commercial power. The lower B+ level is not harmful to the equipment, of course, and other than the decreased power level there is no noticeable effect on over-the-air performance.

design philosophy

Once the choice to produce 110-volt AC power is made, the design can proceed to the block diagram stage. At the start several measures can be taken to reduce the impact of the waveform distortion mechanism mentioned above. The first thing is to minimize the output impedance by selecting a relatively high alternator operating voltage.

The alternator windings normally produce a peak output after rectification of about 15 volts. That voltage is the vector sum of two sine waves phased 120 degrees apart and corresponds to an AC level before rectification of 10 volts RMS. The resistance of the stator winding is 0.286 ohm, so the transformation of 10 volts into 110 volts will create an overall

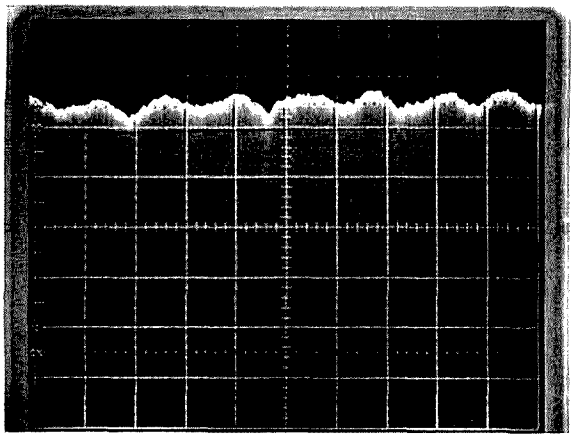


fig. 3. System response to an input disturbance. This 2 to 3 volt (rms) ripple on top of the AC envelope is all that's left of a 30 percent input power variation caused by pedal pumping. Scales are 10 volts and 0.2 second per division. Flywheel energy storage can filter the operator's power strokes enough to mask this problem, but that effect was minimized for this photograph by placing a 100-watt resistive load on the system's output.

output impedance of 34.6 ohms (0.286 ohm times the square of the transformation ratio). Operating the alternator instead at an output level of 12.6 volts RMS will lower that impedance to 21.8 ohms. That higher operating voltage will require a bit more field excitation power, but the lower output impedance is worth the expenditure. In fact, it might be worth trying an even higher voltage transformer if one is available. An 18 to 110 volt transformer would lower the output impedance to 10.7 ohms and possibly provide an overall gain in electrical efficiency as well.

The right transformer can, by lowering the output impedance, reduce the amount of distortion caused by a pulsed current (rectifier) load. The load can be protected against the harmful effects of any remaining distortion by designing the regulator to stabilize the RMS value of the output voltage.

The regulator could be designed to control the peak level of the output voltage, but in that case any distortion will cause the RMS output to increase as the load increases. That's because, as mentioned earlier, large pulse-type load currents will cause the output wave form to degenerate into a square wave. A sine wave with a peak level of 155 volts has an RMS level of 110 volts, but a symmetrical square wave with that same peak voltage has an RMS level of 155 volts.

Before the RMS level can be controlled, it must be measured, and that means an RMS detector is necessary. True RMS converters can be quite complex and expensive, but for this application there's an easy alternative. The high voltage waveform is full-wave rectified and the average DC value of that output is extracted with a suitable low-pass filter. The resultant voltage is not quite the same as an RMS measurement, but is close enough to do the job.

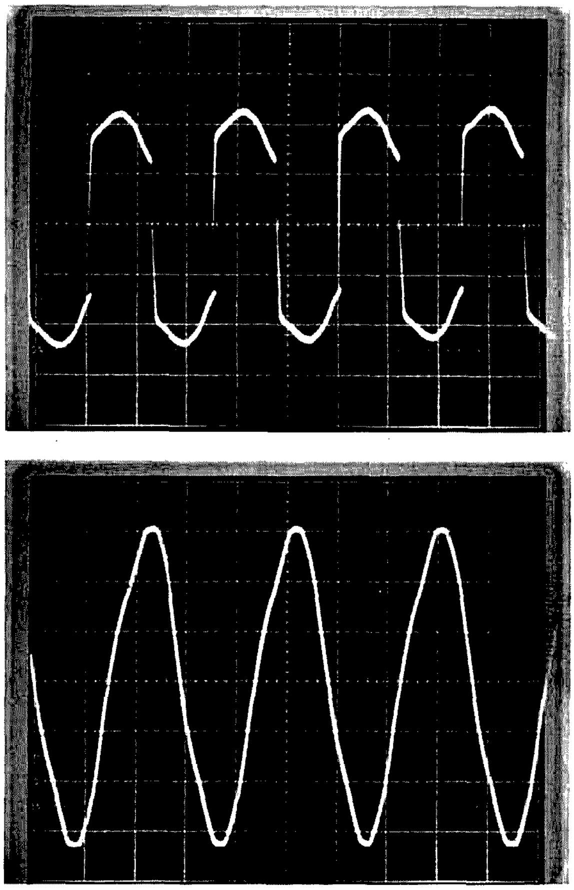
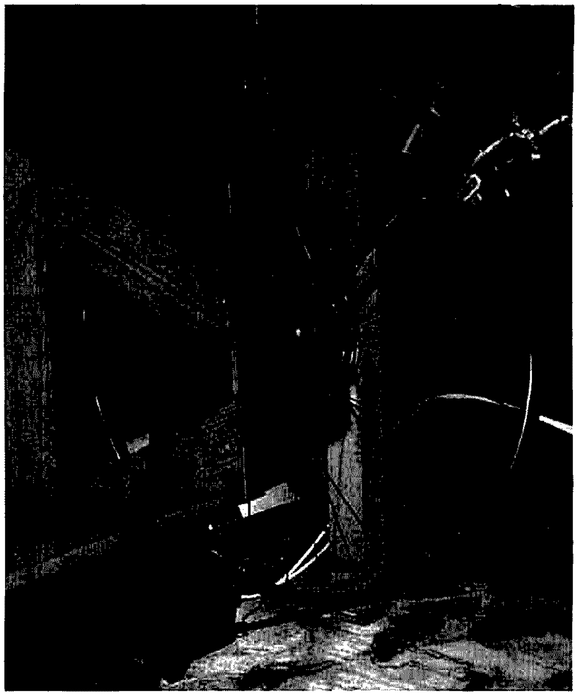


fig. 4. Waveform distortion caused by power supply loads. Scales for both photographs are 50 volts and 1 ms per division. Top photo shows the output AC waveform when the system is delivering 75 watts directly to a resistive load. In bottom photo, that load has been replaced with a transformer-rectifier-filter power supply, which is delivering the same power in DC form to a resistor. As discussed in the text, the peak output level has decreased but the RMS voltage is holding nearly constant.

Another consideration in the design of the regulator circuit is loop stabilization. As with any feedback loop, it's necessary to select components carefully in order to realize the desired response to a disturbance input. When, for example, a sudden 100-watt load is applied, it's expected that the output level will immediately drop about 20 volts (roughly 1 amp flowing through the 24-ohm output impedance). The regulator should then increase the field current so the output voltage rapidly and smoothly returns to 110 volts. In addition to being quick, this recovery should not involve any significant overshoot or ringing.

circuit design

The circuit diagram of the bicycle-powered system



A length of ordinary cycle chain drives a ten-speed sprocket cluster mounted on the alternator shaft. The rear bike rim is wrapped in black friction tape and drives the chain like a large pulley. Crude perhaps, but the drive system is also inexpensive and mechanically efficient. (*Photo courtesy W1GSL.*)

is shown in **fig. 6**. The alternator, shown inside the dashed rectangle, has been rewired slightly to optimize power use. This modification, as detailed in **fig. 5B**, separates one leg of the original Y winding to provide a source of low-voltage DC power for field excitation. Four of the original six alternator diodes are used for this purpose, while the other two are disconnected. The nominal output of this supply is about 10 volts, and that's a good level for Q4's collector supply, since it leaves only 3 or 4 volts to be dropped as waste across that power transistor.

transformer

The remaining two legs of the Y are brought out to drive the primary of the step-up transformer. That AC output is also connected to a bridge rectifier which provides operating potential for the low-level regulator stages. Since those stages don't draw much current, there's no power penalty in running them with this higher voltage, and the extra dynamic range is beneficial. With the regulator powered in this manner, it's possible to run the AC output all the way down to 50 volts before control is lost. The system isn't normally operated at that level, but having that much dynamic range improves the circuit's smoothness during start-up and transient conditions.

The main 12.6 to 110 volt step-up transformer was salvaged from an old black and white television chassis. Any similar transformer, or combination of transformers, having a 12.6 volt winding rated at 8 to 10 amperes will work perfectly well. The overall transformation ratio isn't critical, but primary voltages lower than 12 should be avoided for efficiency reasons.

regulator

The regulator samples the 110-volt output of the system with a small 1:1 transformer. This transformer drives a bridge rectifier through a series circuit made up of various resistors and neon bulbs. This circuit, along with the RC filter that follows the bridge, makes up the RMS (well, not quite RMS, but let's call it that anyway) converter. The idea is that the large effective value of series resistance between the transformer and bridge causes that combination to drive the capacitor like a current source. The capacitor charging and discharging rates are then very nearly equal and determined by the 16.6k load resistance. As a result, the capacitor voltage is free to follow the average value of the rectified voltage instead of its peak.

The neon bulbs are an overvoltage protective feature. Should the AC voltage exceed their firing level, they provide a low-impedance path that quickly charges up the filter capacitor, thus rapidly shutting off the field current. The resistors in parallel with the bulbs are carefully chosen unequal values which help minimize the difference between firing and sustaining voltages. This is about 15 volts per bulb, so the sequential firing encouraged by the resistor choice shaves 30 volts from the total trigger level. That sets the protective threshold at 190 volts peak or 135 volts RMS for a sine wave. Putting the resistors between the transformer and bridge lowers the inverse voltage seen across the diodes so that inexpensive small signal types may be used.

control system

A portion of the filtered RMS output is sampled with the 10K potentiometer and applied to one input of a differential amplifier. A 5.1-volt stabilized reference is applied to the other input, and the difference is amplified with a gain of 15.

The output at the collector of Q2 drives a common emitter stage, Q3, whose gain is partially determined by the impedance of the alternator field winding. It's a little unusual to have the gain of one stage influenced by that of another, but in this case the incomplete buffering is intended as part of the loop compensation. At currents in the 1-amp range the beta of power transistor Q4 is about 40, so the 5 ohm, 0.6 henry impedance of the field winding appears at Q3's collector as a 200-ohm resistor in series with 24 henrys. That impedance combines with the 3.3K collector resistor

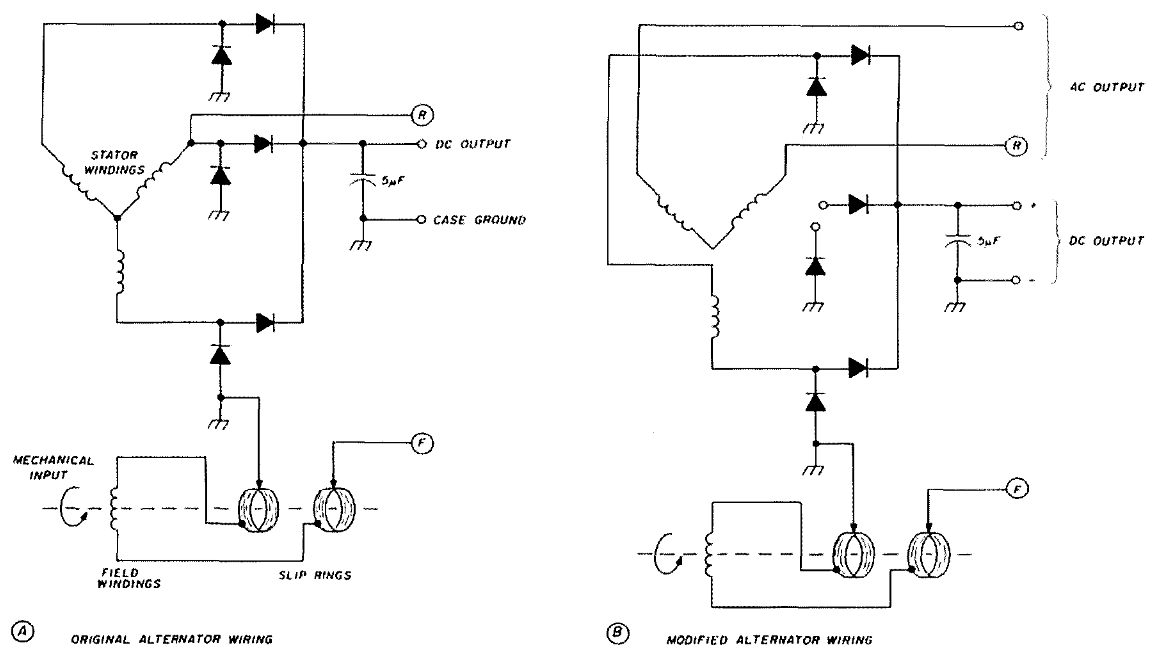
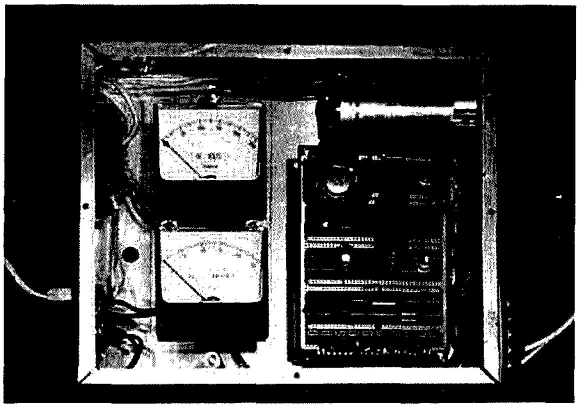


fig. 5. Electrical changes to the internal wiring of the Delco-Remy automotive alternator.



The controller electronics are mounted in and around a standard chassis. Metering is provided for alternator field current and AC output voltage. A set of four flashlight batteries, mounted out of sight behind the circuit board, provides field current through a push button when the system is first started. (*Photo courtesy W1GSL.*)

to create the lead-lag compensation necessary to stabilize the system.

The output transistor, Q4, operates as an emitter follower, and as always with emitter followers, some care is needed to prevent high-frequency oscillations. No instabilities were observed with this circuit, but some high-frequency loading (the series 0.001μF and 1K ohms) was added in any case. The 10μF capacitor is a termination for the AC voltage generated in the field winding by the stator current (back EMF).

One last circuit feature is startup field current provided by 4 D-size batteries through a momentary contact push-button switch and diode. When the setup is first started, this button is pushed and the alternator brought up to speed. The output generated by the battery current is enough to power the regulator, which can then bootstrap the system on up to 110 volts. When that happens, the diode in the battery circuit back-biases so the regulator can control the output level, and of course the operator can release the pushbutton.

mechanical design

Most of the mechanical details are visible in the photographs. The old and inexpensive 10-speed bike that drives the alternator is mounted at its rear axle to a large and stable base made from scrap lumber. The front wheel is removed and replaced with a wooden support. (While not strictly necessary, this adds stability to the setup by discouraging rotation of the front fork.) The alternator itself is mounted much as it would be in a car. Simple but sturdy aluminum brackets accept a mounting bolt that can be loosened to allow tension adjustments of the drive chain.

The alternator is chain-driven from the rear rim of the bicycle. The tire and tube were removed from the rear wheel, and the rim was trued up as much as possible with a spoke wrench. Any protruding spoke ends were filed flat and the inside of the rim was wrapped with several layers of black cloth friction tape. Mean-

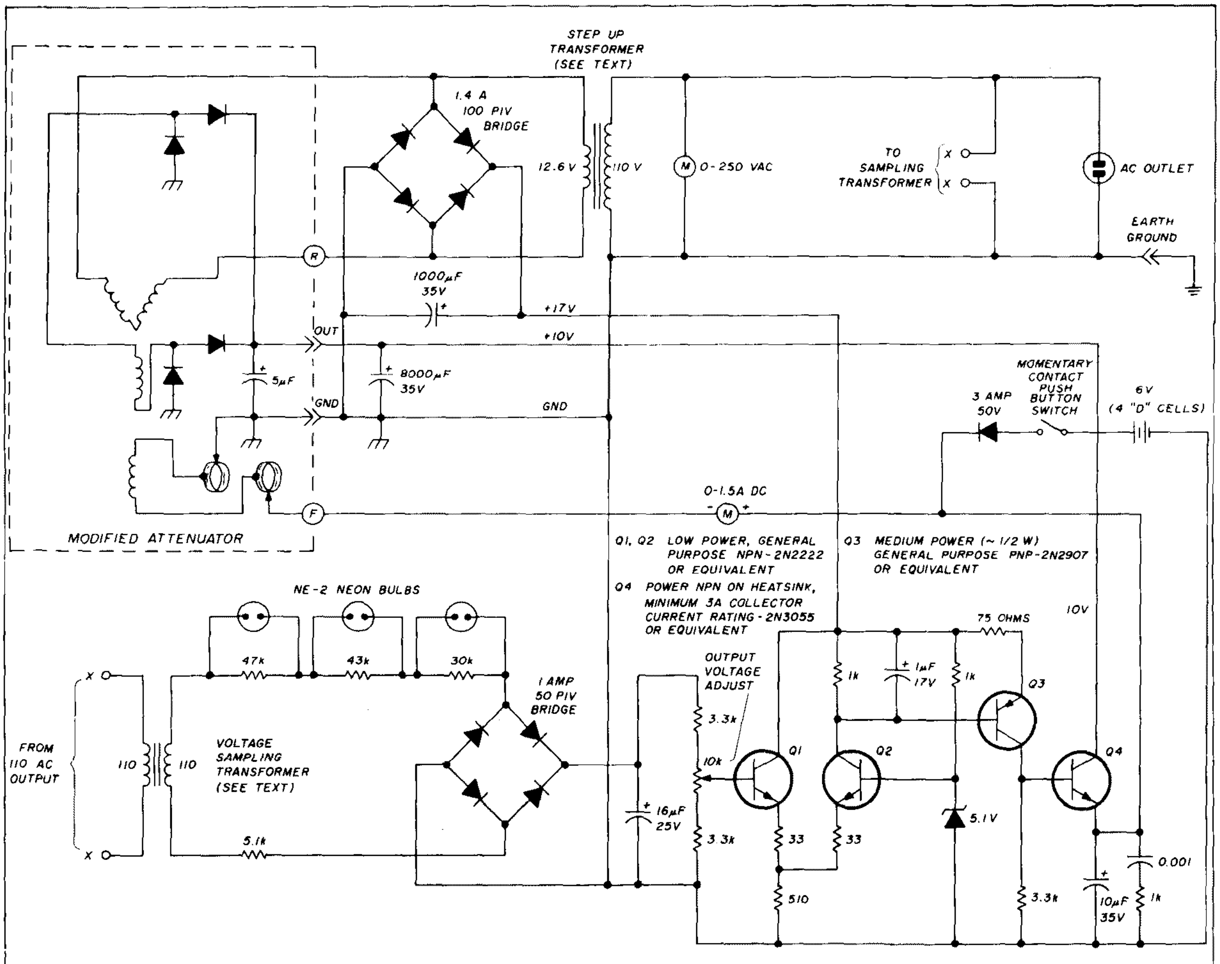


fig. 6. Schematic diagram of the bicycle-driven 110-volt power supply.

while, the alternator shaft had been fitted with a standard 10-speed rear sprocket cluster, so the bicycle rim and alternator could be mechanically coupled with a length of standard 10-speed bicycle chain.

This drive system may be simple, but it's also remarkably efficient, trouble-free, and inexpensive. The chain is friction-driven by the wheel rim, but the contact length is long (about 5 feet) and there is no noticeable slippage. The roller chain is quite flexible and has a smooth but positive grip on the alternator drive sprocket. The only thing at all fussy about the drive is that the bike rim must be reasonably circular since any radial runout will cause a variation in chain tension as the wheel revolves.

The drive chain used is custom-fitted to the mechanical setup by combining portions from two standard length chains. Ten-speed cycle chains have no master link and are instead adjusted with a special tool that disassembles links by pushing out the rivets. Available for about $5 in most cycle shops, that inexpensive gadget makes it possible to easily adjust the length of the drive chain.

The 10-speed sprocket cluster, as shown in **fig. 7**, is held on the alternator shaft with much of the original hardware. A simple spacer turned from scrap metal centers the cluster on the shaft and provides a flat surface for the clamp nut to seat against.

electrical construction

There are no sensitive areas in the regulator circuit, so layout and construction can really take any convenient form. The controller shown in the photographs was built in and on a 10 × 12 × 3 inch (25 × 30 × 7 cm) aluminum chassis. The leads running to the alternator (No. 8 AWG wire was used for the high-current AC wiring) connect to a barrier terminal strip on the right side of this chassis. The AC output leaves through a standard AC receptacle mounted on the left. Directly above the AC outlet is the step-up transformer. A small 1:1 isolation transformer, mounted on

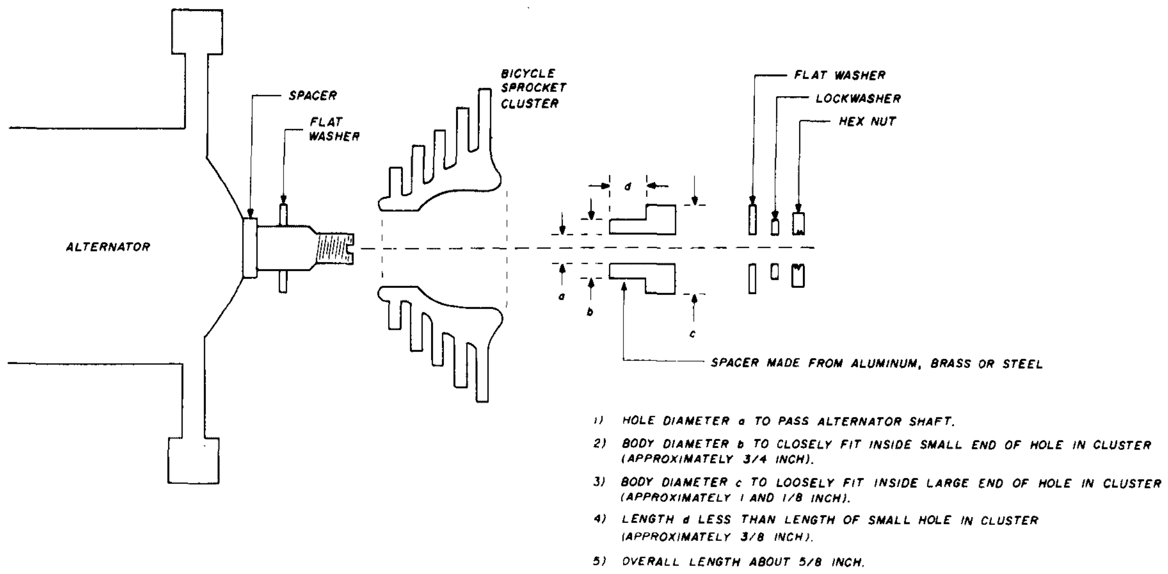


fig. 7. Method of mounting the sprocket cluster to the alternator shaft.

the right side of the box, samples the output; aside from being expensive if purchased new, it's bigger than necessary. A pair of very small filament transformers connected back-to-back would work just as well since the RMS converter circuit draws only a miniscule amount of power.

The smaller components of the regulator circuit are mounted on a standard 4.5 × 6 inch (11 × 15 cm) circuit board mounted inside the chassis. The plug-in feature isn't necessary, but was convenient to use during the development stage of the project. The 4 D-sized cells used to provide startup current are mounted out of view behind the circuit board.

In keeping with standard safety practice, one side of the AC output is tied to the chassis, which is in turn properly grounded. This step is necessary to prevent stray coupling and leakage paths from creating a large potential difference between the AC outlet and radio ground.

checkout and operation

Operation of this system is simplicity itself, but initial checkout should proceed with care. There is a surprising amount of power available here, and some expensive radio equipment is likely to be plugged into it. *Don't take any chances with your transceiver until you're sure the regulator is functioning properly!*

A lot of testing can be done without any load at all on the system. First set the voltage control potentiometer to mid-position. Press the field excitation pushbutton and start pedaling. The AC output voltage will slowly increase to 50 volts or so, then jump upscale to its final value. This level can then be set between 110 and 115 volts with the pot.

The neon bulbs may flash briefly when the output voltage first swings upscale. That's because the filter capacitors in the regulator power supplies take a moment to charge up. Until they do, the regulator isn't fully operative and the output AC level can build up rapidly. A worst-case example of this effect is shown in **fig. 2A**. The large overshoot in this photo resulted from jerking the alternator up to high speed as rapidly as possible. The size of the overshoot can be reduced a bit by smoother pedaling, but it always makes sense to have the output voltage up and stable at 110 volts before an electrical load is applied.

If everything seems satisfactory at this point, start loading the output. Begin by trying light bulbs of different wattage so you can develop an appreciation of how much effort is required to produce so many watts of power. An oscilloscope can be used to check how the system responds to suddenly applied loads. Light bulbs draw quite a bit of inrush current when first switched on, so they don't make a particularly good load for transient analysis. A soldering gun is better for this purpose because it represents an almost purely resistive load and can be switched on and off easily.

Other interesting loads include things like old five-tube table radios and small television sets. *Repeat: Only when you're convinced the circuit is working smoothly and reliably should you plug in a valuable*

transceiver. The circuit of **fig. 6** is a mature design and works quite well, but I'm sensitive to all that can go wrong because I burned out a large number of 100-watt light bulbs in the early stages of this project!

performance

The measured electrical performance of the system is detailed in **figs. 1** through **4**. No effort was made to measure mechanical performance due to the lack of both suitable instrumentation and a data base to use as a yardstick.

Electrical measurements are much easier to make, and **fig. 1** shows two of the most important: steady-state regulation and overall electrical efficiency. To make these curves the system was pedaled at a constant rate while connected to several 110-volt light bulbs through a calibrated AC wattmeter and Variac. The output voltages and currents were then measured for a range of different load levels. The curve showing overall electrical efficiency is a measure of how much power was generated as compared to how much went to the load. The lost power was used either for field excitation or dissipated as waste in the alternator and transformer resistances.

The transient response photographs in **fig. 2** show how well the regulator responds in dynamic situations. The first picture shows the output AC envelope during the system power-up transient. The second part of **fig. 2** shows how the regulator handles the sudden application and removal of a 75-watt resistive load. It's difficult to maintain a constant pedal rate in the face of that load change, so the picture unfortunately shows the combined response of both the regulator and operator. Even with that complication the loop looks pretty good.

There is some flywheel energy storage in the moving components (most noticeably the high-speed chain drive), but when the electrical loading is heavy, that filter isn't enough to smooth the two-per-cycle power strokes coming into the system. **Figure 3** shows how well the regulator can hold the output level constant under those conditions.

Figure 4 shows the distortion effects in the output waveform caused by a power-supply load. The first trace was taken with 75 watts of resistive loading on the system. In **fig 4B** this was replaced with 75 watts of DC power delivered through a full-wave bridge rectifier to an RC load.

conclusions and suggestions

The system described has been in use for over nine months. I try to operate for at least half an hour every other day, so that nine months represents a lot of air time and a reasonable amount of physical conditioning. My transceiver draws 40 watts of power on receive and 80 watts of steady-state bias power on transmit. I can power the radio itself for well over an hour without tiring, so for exercise purposes the system is also loaded with a 40-watt light bulb.

With that loading the system produces 130 or 140 watts during transmissions and the output ripple due to pedal pumping becomes bothersome. As an easy cure, some 4 or 5 pounds (1.8-2.3kg) of flywheel weighting was added to the rim of the bicycle's rear wheel. That provides enough energy storage to operate the setup for several seconds, so even at high output levels the effect of pedal pumping is completely removed.

The only other modification under consideration at this time is changing the step-up transformer to an 18-volt unit. This would cut the system output impedance in half and so improve regulation, reduce waveform distortion, and increase overall efficiency in the 100 watt output range.

acknowledgement

Special mention must be made of the efforts of W1GSL, my trusty (and trusting) photographer. Not only does he make house calls, but he was willing to give the system a trial spin in his street clothes before getting down to photography.

ham radio

the new industrial revolution: packet radio and local area networking

"I see no reason why intelligence may not be transmitted instantaneously by electricity."

— Samuel Morse, 1832

Because the traditional methods of point-to-point communications — CW and RTTY — relied on relatively low data rates, conventional systems have been slow to take advantage of available technologies and the bandwidths they offer. Yet the widespread use of digital computers has created an urgent demand for faster data transfer (burst communication) requiring large bandwidths giving a totally new meaning to Morse's words.

Now that personal computers have taken the place of comparatively "dumb" terminals, used to access large central computers (CPUs) via low-speed telephone lines, communication between and coordination of individual computers has presented a new and difficult challenge to system planners. Because there are differences between various computers, they have difficulty "talking" to each other, and problems are created whenever they are combined in a common network. In order to solve this problem, a new science, called LAN or *L*ocal *A*rea *N*etworking, was created, which employs packet-like communications to transmit data on dedicated wire (pairs), coaxial cables, fiber optics or "through the air" (see **fig. 1**).

The need for wideband communications has also affected traditional analog modulation schemes, which are now being replaced by digital ones. The search is on for techniques that will enable diverse types of electrical information — including voice and video signals — to be combined and transmitted. At the same time new higher speed analog-to-digital converters are replacing conventional processing techniques not only in RF processors but in large distributed networks and private branch exchanges (PBX). In addition, compressed full-motion digital television signals, including digital voice and computer information, can be distributed over networks for videoconferencing.

This phenomenon is not a science fiction fantasy, but rather a reality of the new industrial revolution, in which information is managed through automatic communications between and among "intelligent" machines.

As an example, consider a corporation with several manufacturing facilities in widely separated locations. Each factory acts as an independent unit with its own offices and production facilities. To operate effectively and maintain the competitive edge, each unit needs to have, at hand, easy access to the latest technology and techniques of information processing and communications. The answer to this need is a series of interconnected local area networks deisgned to instantly respond *electronically* rather than on paper.

communication path determines technique used

Most data communications over paths shorter than one hundred meters employ (hard-wired) parallel busses (see **figs. 2** and **3**.) Although this relatively fast method of data transfer is fully compatible with computer I/O's, use of busses over greater distances becomes impractical. Consequently, local area networks use serial digital packet formats at rates in the 50 Mbit/s range.* In communications this is referred to

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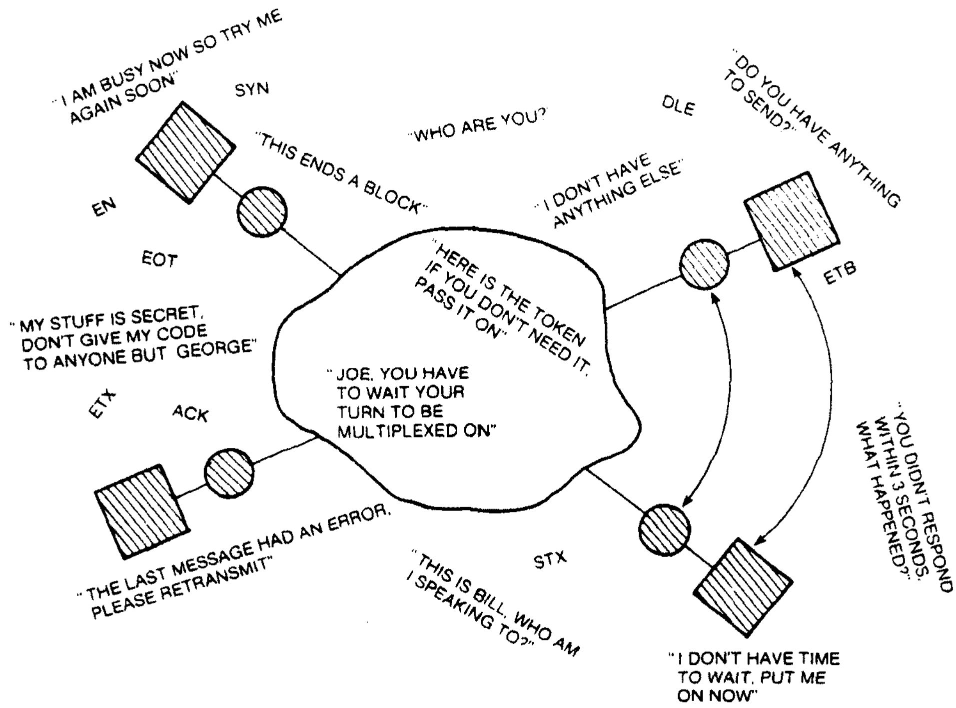


fig. 1. Some protocol requirements for local area networks.

as base-band information because it is the basic format that every device will eventually generate and accept for communications. Base-band can be used directly in point-to-point communications between data devices such as universal asynchronous receiver/transmitters (UART). However, without complex protocols and equipment, only two devices can communicate over the given medium at a given moment. For communication among more than two devices (e.g., computers), other methods must be used.

One method of communicating between several digital base-band devices over the same pair of wires is by multiplexing them synchronously at both ends. Known as time division multiplexing (TDM), this represents an early form of local area networks, as shown in **fig. 4**. TDM uses permanently dedicated time slots that can sometimes be "empty" of information; *statistical* time-division multiplexers (STATDM) eliminates this problem by dynamically allocating the time slots according to the activity on the channels.

- ***Computer Bus:***
  1-10 cm to 100 meters.
- ***Local Area Nets:***
  Several meters to several kilometers.
- ***Local Distribution Nets:***
  A few kilometers to approximately 100 kilometers.
- ***Long Haul Nets:***
  Hundreds of kilometers to thousands of kilometers.

fig. 2. Categorization of networks by geography.

*Packet is a block of data handled by a network in a well-defined format including a header (opening ID) and having a maximum size and data field. Consequently, a message may have to be carried as several packets.

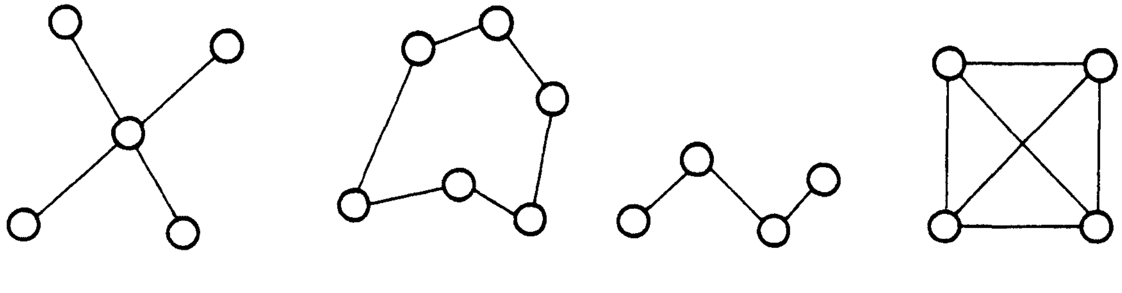


fig. 3. Types of local area networks.

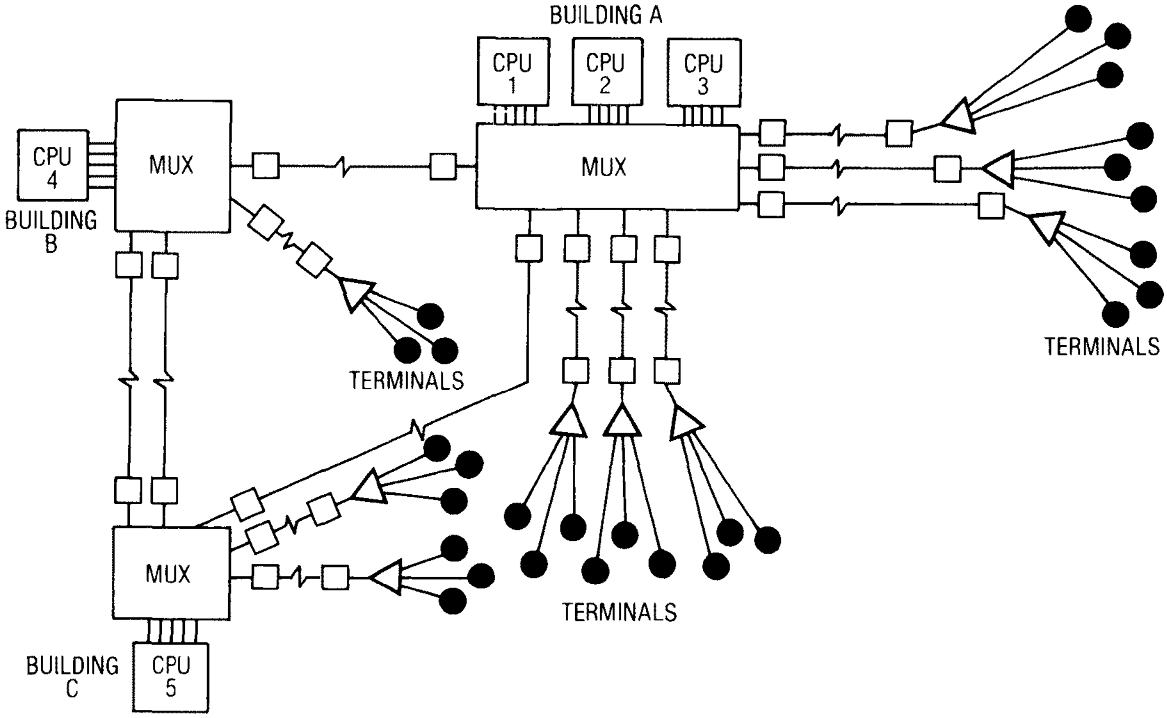


fig. 4. Block diagram of a campus local area network using multiplexers.

Other simple local area networks use an intelligent central switching facility in a star configuration that resembles conventional telephone PBX technology. By using base-band directly on wires and without multiplexing or central switching, we occupy 100 percent of that medium's bandwidth. (Also known as the percentage bandwidth factor.)

When base-band signals are translated to radio or optical frequencies, it becomes possible to pack more base-band channels into one single frequency, as shown in **fig. 5**. The higher the frequency, the more base-band *channels* can be transmitted.

Frequency division multiplexing, designated as the broadband approach to local area networks, can be further subdivided according to some interesting protocols, as shown in **fig. 6**.

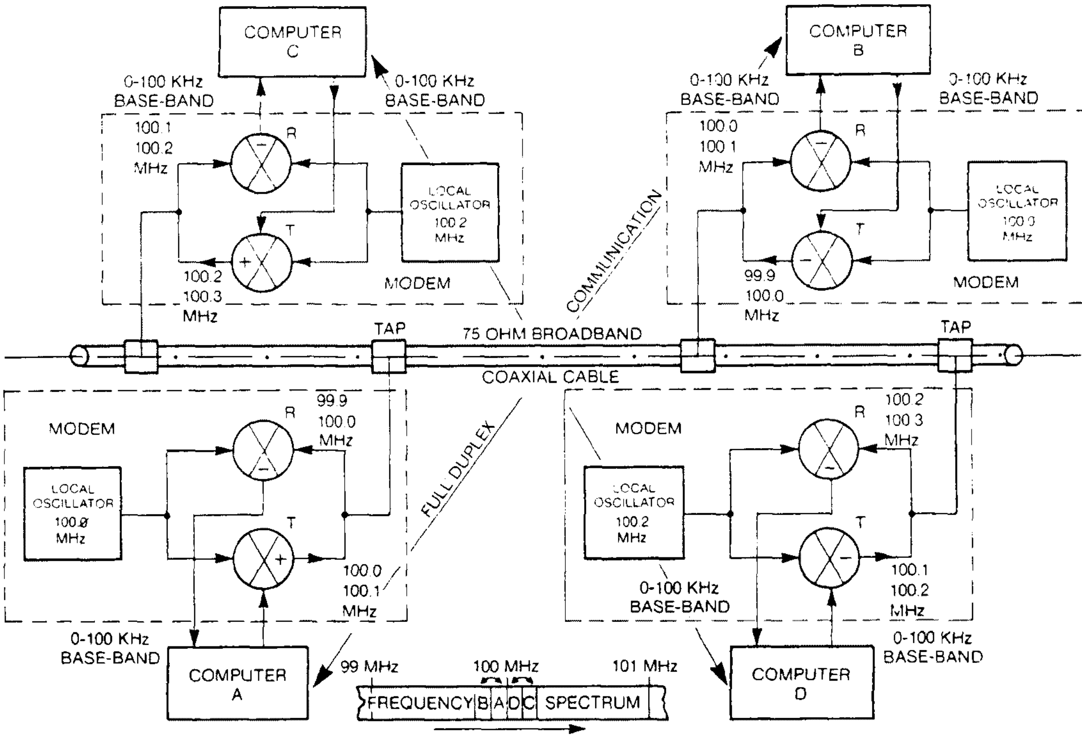


fig. 5. The concept of percentage bandwidth shows that the best medium utilization occurs when many base-bands are modulated on top of much higher frequencies as in frequency division multiplexing (FDM) techniques using radio-like processing.

listen while talking

In an effort to find better ways of utilizing the spectrum *without* switching and/or multiplexing new packet techniques using radio and coaxial cable networks have recently been developed and implemented. By borrowing ideas from real-life situations, designers *have* applied the characteristics of human interaction to communications between machines.

Imagine a business meeting in which several dynamic men and women exchange information and make decisions in one *room*. The process is rapid and contains several important elements common to all the speakers:

- A leading address to one or more participants preceeds any conveyed message.
- A message is conveyed.
- An end remark is usually followed by an invitation to answer.

If two persons begin to speak simultaneously their messages "collide," and one or both stops talking and tries again later. (In earlier packet radio, neither party would have "known" that a collision had occurred until they noticed that an acknowledgement had not been received at the end of the transmission.)

This relatively simple concept is the basis for a type of local area networking known as *Ethernet*. Ethernet allows the transmitting station to listen in on the "ether" while transmitting a packet, in a method known as "listen while talking." This method is similar to the situation that occurred in the hypothetical business meeting described above. It allows for any

collisions caused by another station that had decided to transmit to be immediately detected, which in turn causes the transmitter to cease transmission and restart at a later time. Ethernet can be used either at base-band over wire pairs and/or coaxial cables, or over channelized broad-band busses in CATV-like networks equipped with special modems as shown in **fig. 7**.

The concept was developed as a consequence of problems with simpler packet radio techniques that do not necessarily allow for the collision detection characteristic of Ethernet. These technologies evolved from the well-known ALOHA network. Developed at the University of Hawaii in 1970, the ALOHA system was essentially the first practical packet radio network, and covered the Hawaiian Islands, as shown in **fig. 8**.

The purpose of the ALOHANET was to provide inexpensive, statewide user-to-central computer communication for several hundred terminal users who were experiencing poor telephone modem connections at the time. Unlike previous networks, which used node-to-node communications, ALOHA network communication was accomplished in a radio broadcast mode, in which each station was heard and addressed by the central packet station known as the "Menehune", located in Honolulu, via two 100-kHz random-access channels at 407.35 MHz and 413.475 MHz. Although the system did not allow for direct user-to-user communications, information was nevertheless transmitted by transferring data to the central packet station and then forwarding it, after processing, to the destination user.

The packet transmission data rate was 9600 baud, with packets consisting of a 32-bit header, a 16-bit header parity check field, and up to 80 bytes of data followed by a 16-bit data parity check field. The maximum size packets were, therefore, 704 bits in length. Each took about 73 milliseconds to transmit; the entire network's propagation delay time was therefore, negligible in comparison to other systems. The random-access user-to-computer (407.350 MHz) channel allowed for packet headers containing user addresses to be identified by the Menehune central station. One natural consequence of the random nature of transmission was the probability of packet overlapping and collision. However, frequency division multiplexing was not considered necessary because of the burst nature of the computer traffic and the lack of additional radio channels. The collided packets were rejected by the Menehune station, and that fact was made known to the respective transmitting terminal by the absence of an acknowledgement signal on the computer-to-user return channel (413.475 MHz). This channel presented no random collision problem, since only the central processor's transmitter was broadcasting to all the other stations.

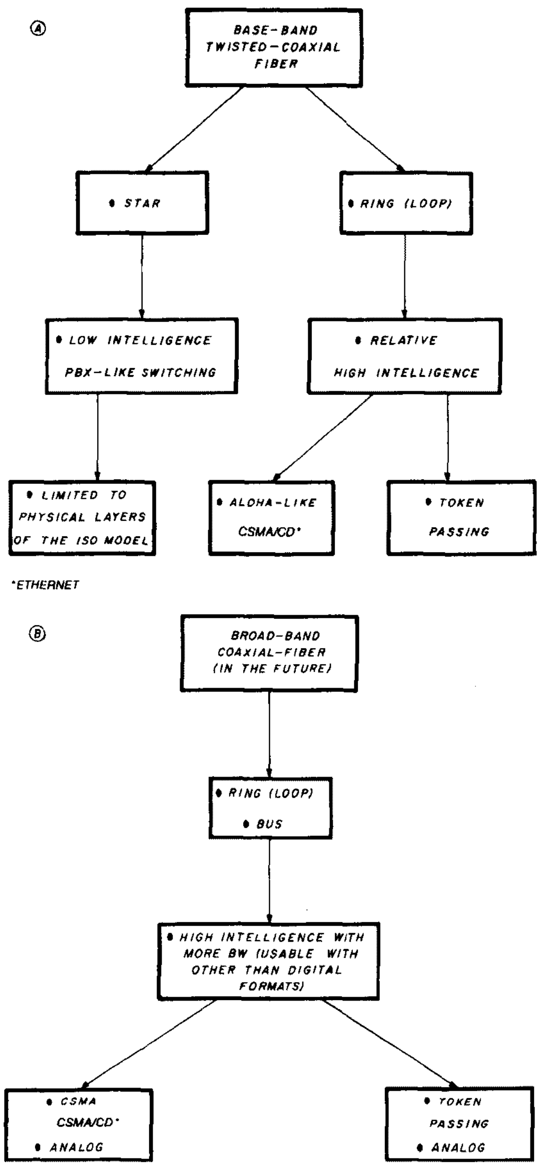


fig. 6. Characterization of networks by frequency domain and protocols; (A) using coaxial cables, (B) using fiber optic cables.

This scheme, known as pure ALOHA, presents some drawbacks as pointed out earlier; because the individual terminals have to wait until no acknowledgement is received, and then re-transmit the packet at random, with no guarantee that a collision will not occur again.

The relatively low throughput of pure ALOHA (see **fig. 9 A** and **B**) arises from the high practical probability of packet collision, which in turn is the result of

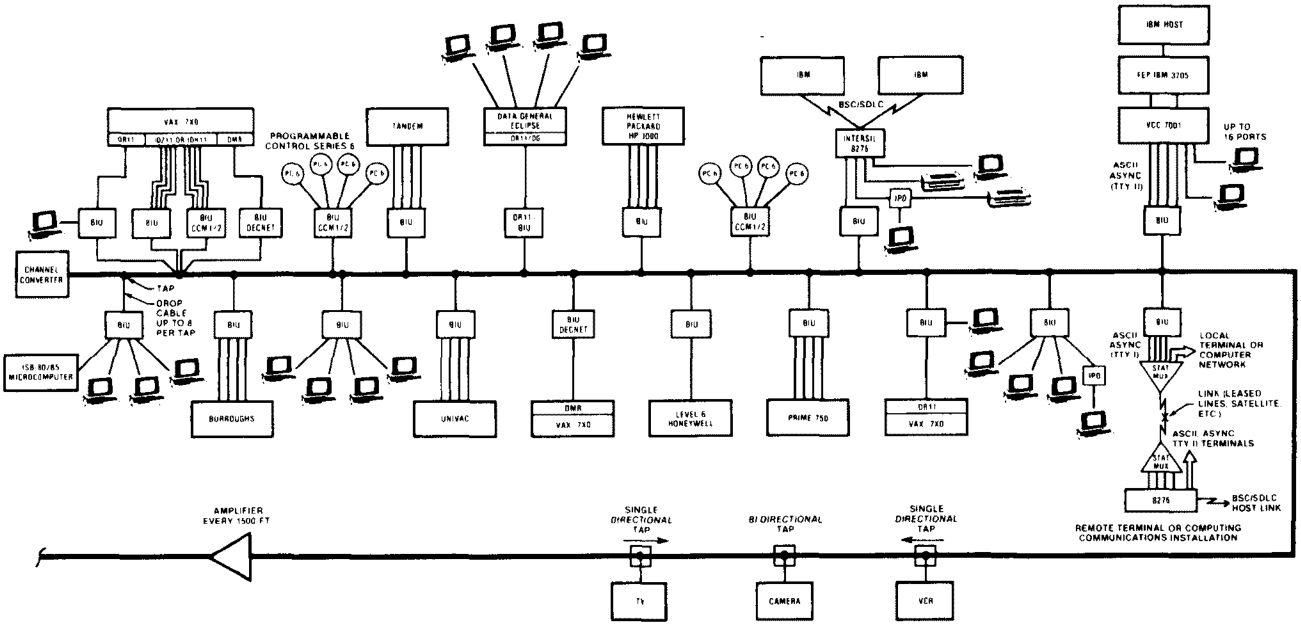


fig. 7. The GE-NET is an example of Ethernet implemented on a broad-band CATV-Network.

the total lack of discipline in transmissions from the terminals.

To correct the situation, various proposals were made to improve the bandwidth utilization. Among them were slotted ALOHA, which restricts and synchronizes terminal transmissions for better throughput and the various carrier sense multiple access (CSMA) techniques presently used.

The present AX.25 protocol used in Amateur packet-radio networks is superior to the ALOHA protocol because it listens before it transmits; however, the characteristic "listen while talk" feature of Ethernet is not available on these networks. The more advanced Ethernet concept was pioneered by the Xerox Corporation in November, 1978, as part of a proposed nationwide network known as the Xerox Telecommunications Network (XTEN).

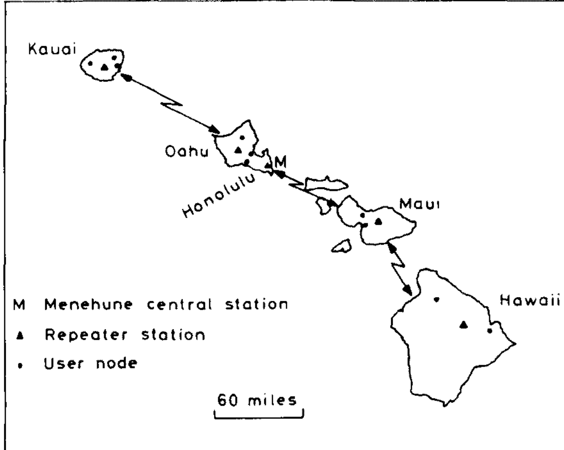


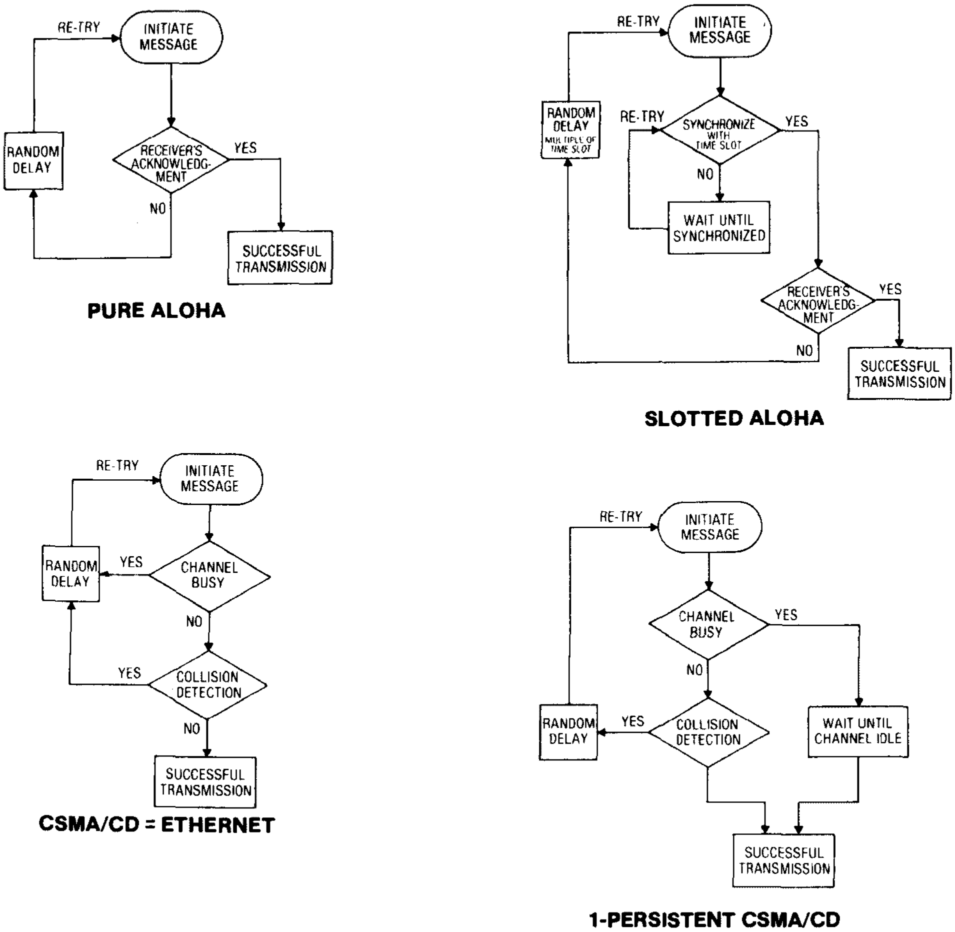
fig. 8. Map of the ALOHA network.

satellites would provide the link

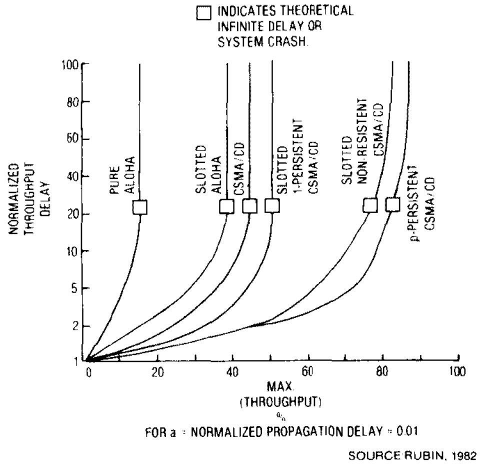
The overall XTEN approach required satellite communications to link together through shared earth stations located in all the major cities around the country. Microwave radio, and/or infrared fiber links were envisioned to connect the local users to the earth stations, and finally, individual buildings and/or campuses were to be connected through a 3-Mbit/s packet network.

In its modern form, Ethernet is a base-band local area network allowing a theoretical rate of 10-Mbit/s over coaxial cable. It can connect as many as 1024 stations over a span of 7600 feet. Each station is designated by a 48-bit code and collision detection (CSMA/CD) is accomplished through the previously described techniques. In spite of the many improvements since the ALOHA days, the opponents of Ethernet still maintain that when network traffic is especially heavy, real-time applications such as those on aircraft or in automated factories could fail to be completed within the limits of their prescribed times, resulting in long collision recovery time, and even the risk of catastrophic failure.

Consequently, new methods have been developed to provide more discipline in accessing networks. One



A



PROTOCOL	MAX. THROUGHPUT (%)
PURE ALOHA	18.4
SLOTTED ALOHA	36.8
CSMA/CD	40.0
1-PERSISTENT CMSA/CD	52.9
SLOTTED 1-p CSMA/CD	53.1
0.1-p CSMA/CD	79.1
NONPERSISTENT CSMA/CD	81.5
0.03-p CSMA/CD	82.7
SLOTTED NONPERSISTENT CSMA	85.7
IDEAL SCHEDULING	100

B

fig. 9. (A) Variations of the ALOHA concept, and their functional flow diagram, (B) system crash is indicated by the maximum throughput delay for the different methods shown.

such method is the newly introduced *token-passing* technique, which consists of passing an accessing token from terminal to terminal, which in turn provides a well-ordered method of accessing the network for the participating users. The relatively complicated error recovery algorithms of token-passing make this approach difficult to implement, however. (Most base-band, token-passing networks can also be implemented in broad-band.)

A compromise between the CSMA and *token-passing* is now being examined, with products to follow in the near future.

applying what we've learned

Shown in **fig. 10** is a radio transceiver with a built-in computer terminal; it could also be considered a computer terminal with a built-in radio transceiver. Behind its friendly look lies a packet radio network of unprecedented complexity.

Imagine a nationwide dispatching operation involving thousands of roving service or supply vans, with drivers. Obviously such an operation would require foolproof, effective communications and positive inventory control of all parts used in the field (with automatic reordering). A combination of radio paging, telephone dispatching, and paperwork would be necessary.

In such an application increased productivity and elimination of delays associated with the paperwork means improved communications. Equipped with this portable radio terminal, otherwise known as a PCX, a fleet can reliably access a distributed database network that keeps track of all inventories even as they are being continously modified from the field by all other portable radio terminals. This — and dispatching and ordering functions as well — happens in real time, in all locations, and without the inevitable delays associated with paperwork.

These developments are expected to open new doors in professional and Amateur packet radio activities, with products such as hand-held transceivers expressively made for data networking applications. Networks using satellite packet gateways will allow electronic mail to be carried from one side of the world to the other, and facilitate the handling of emergency traffic on a worldwide scale.

what we can expect

Today's personal computer is a small first step in the eventual connection of diverse communication technologies in terms of interactive thinking of machines. This new science has been named *telematique* or *telecybernetics*.

Theoretically, the ideal local area network should allow any computer or office/factory device to communicate with any other device within its geographical

fig. 10. The portable radio data terminal (PCX) can provide interactive data communications with large data bases geographically dispersed. Thousands of such devices can modify inventories in real time, while roving around cities as part of area-wide packet networks operating in the 800-MHz business band. The local networks are in turn connected with a nationwide data network which provides updating of information to all other PCXs located elsewhere on the network. *(Photo courtesy of Motorola Inc.)*

boundaries. However, current local networks are limited in that they perform only limited functions in specific settings. This approach may or may not survive the present technological revolution. Such is the case with some inexpensive base-band systems that can work with only certain brands of computers. *The problem is essentially one of standardization.* The age of the throwaway computer is here; the average life time of a design appears not to exceed two years. And although the new machines are increasingly powerful, with built-in LAN capabilities, combining them into networks remains one of the most challenging tasks ever encountered.

To help minimize these problems, the International Organization for Standardization (ISO) introduced the Open System Interconnection (OSI) model in 1978, a concept for developing communications between dissimilar devices. The model has since been adopted by the IEEE and is now reflected in the IEEE-802 standard, as shown in **fig. 11**. Upon completion, this standard is expected to ensure compatibility between local area networks with only a minimum amount of work required of the user.

conclusion

Today's local area networks cannot be expected to provide communications between dissimilar devices without additional customized hardware and software interfaces. At the same time, standardization is not expected to resolve the fundamental incompatibility of computers that use different operating systems, languages, and syntaxes. Technical experts seem in

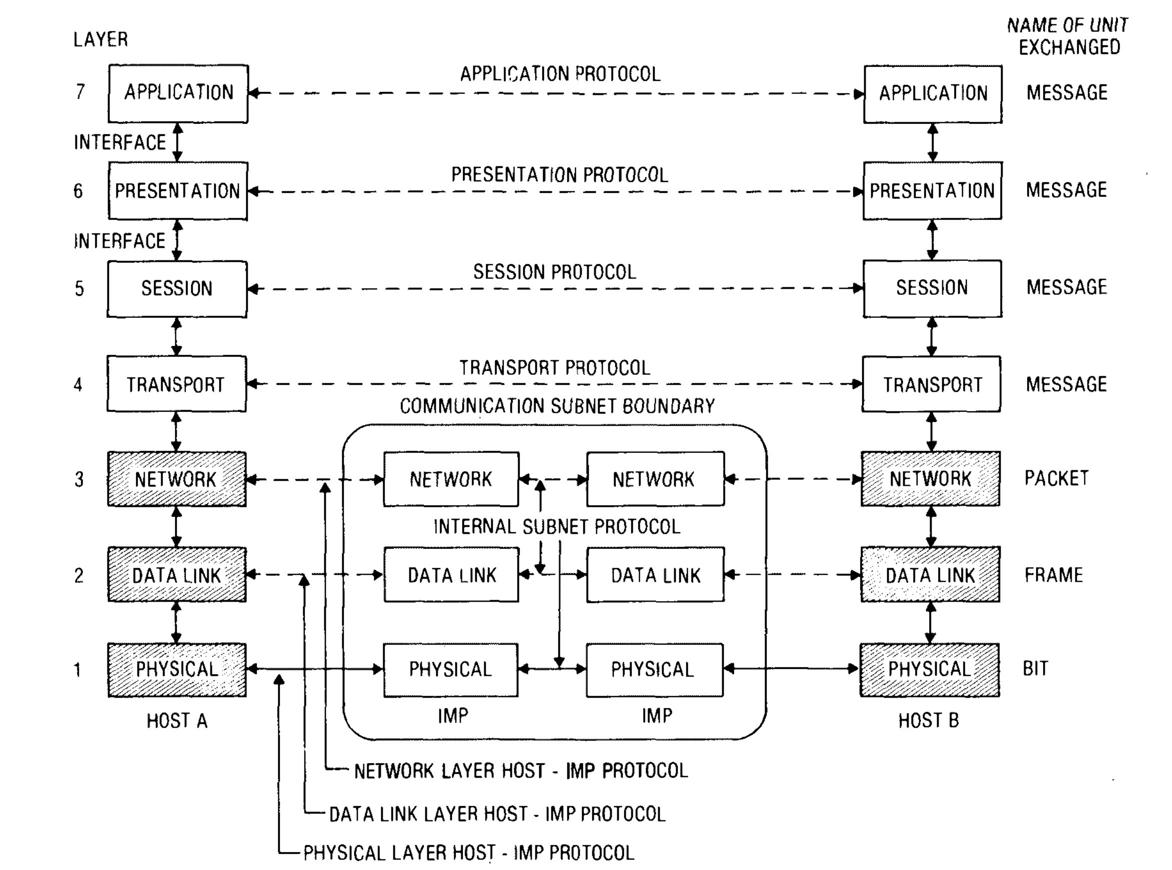


fig. 11. The seven-layer ISO OSI model has been adopted by IEEE's Project 802 Committee and a set of protocols is now available for the bottom layers.

doubt as to whether the processing imposed by the protocols will still permit real-time applications such as digitized voice or compressed video-conferencing — services that have been part of the promised land of LAN. The question still remains whether or not electronic mail belongs within the ISO OSI structure, outside of it as a separate service, or somewhere in between. It is well known that few true distributed database products are available today. Not to be confused with file sharing, these systems involve similar databases that are geographically dispersed. The intent is to provide each LAN containing such a database with updated information in real time. The problems associated with finding, updating, and verifying this information at several locations calls for traffic-intensive communications not yet compatible with the ISO OSI model because of the current lack of "intelligent" software.

The hardware-oriented issues with packet communications seem small when compared with the above issues; progress remains to be made in frequency division multiplexing technologies required for broad-band optical communications and in the design of fiber-optic power splitters (the optical equivalent of RF-power splitters). At light frequencies the process becomes impractical because of temperature instabilities. Consequently, achieving broad-band communications over fiber optics remains a laboratory experiment. Fibers are usually used in base-band modes; applications using Bragg technology are possible, however, in real time broadband-to-base-band converters (patent pending by the author and Honeywell, Inc.).

acknowledgements

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ham radio

polymer film transforms mechanical energy to electrical energy

New material promises a variety of applications, including a simple keyer you can build

Polyvinylidene fluoride (PVF_2) is a flexible polymer film that exhibits both piezoelectric (mechanical-electrical) and pyroelectric (heat-electrical) activity. Unlike the conventional ceramic piezoelectrics found in microphones and record player pickups, PVF_2 film is thin and flexible and can be cut to any desired shape. PVF_2 film has been used in applications ranging from microphones, headphones, and keypads to transducers that reduce the buildup of algae on ships' hulls.

Seeing the potential for its use in both Amateur Radio and non-Amateur related applications, I recently spent $35.00 for a sample package of PVF_2 film and began experimenting. This article relates what I've learned over the course of my experiments.

how it works

To understand PVF_2 film it's necessary to examine its construction. **Figure 1** represents a single sheet of PVF_2 film. Its thickness varies from 0.240 mil (6 μm) to 4.4 mil (110 μm), depending upon the manufacturing process used. The nickel metallization layer is only 0.002 mil (500 Å) thick.

As **fig. 1** shows, the molecules in a PVF_2 sheet are arranged in a zigzag pattern much like the folds of an accordion. Two fluorine atoms and one carbon atom form a molecular dipole, causing a net cigar-shaped charge redistribution within that region. Likewise, two hydrogen atoms and one carbon atom form a molecular dipole with an opposite charge redistribution. Following the pattern, the dipoles are arranged in a head-to-tail fashion throughout the entire length of the material. Collectively, a net electric field polarization exists within the material in the direction indicated. It is the physical movement of these dipoles that cause a voltage to be produced or a physical displacement to occur whenever a voltage is applied. Because voltage and *molecular* motion are *interrelated* in PVF_2 film, the material can be stretched to produce a voltage or a voltage can be applied to the material to cause it to move.

When PVF_2 film is stretched in the direction shown in **fig. 1**, a voltage with the indicated polarity is shown for this particular action. As the material is relaxed, the molecular dipoles return to their original position and a voltage of opposite polarity exists until the material once again stops moving. For PVF_2 film, the voltage output per unit displacement is typically 65 $\times$ 10^7 V/m (volts per meter); thus a molecular displacement of 0.0004 mil (0.01 μm) (a typical value) produces a voltage output of 6.5 volts.

stretching speed determines output voltage

One topic that requires further attention is that of the apparent "rate effect" present in producing voltage from mechanical stress. Although a voltage is produced by physically moving the molecular dipoles, the charge present in the material itself is mobile and redistributes itself to maintain an overall charge neutrality. To produce a substantial voltage output it is necessary to rapidly move the material along its stretch axis by applying a quick mechanical impulse. The voltage output is greatest for a quick, sharp tug (maximum impulse) and lowest for a gradual pull. Also, if the material is relaxed quickly, a sharp, opposite-polarity pulse is generated. The voltage output which results from the mechanical displacement is related not only to the actual displacement but also to the *rate* of displacement. The voltage output is not a steady DC voltage; rather it is a voltage pulse occurring initially with one polarity and then (upon relaxation) the other.

Figure 2 illustrates the effects of these actions. In this case, the material is pulled taut along the stretch direction and is maintained under tension, much like the material stretched over the head of a drum. The initial positively-sloped line at the left hand side of the figure (between points 1 and 2) indicates the maximum impulse region and thus the voltage produced is a maximum. Between points 2 and 3 the force is con-

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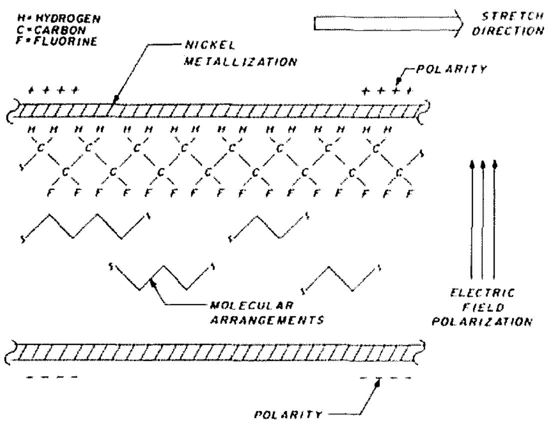


fig. 1. Cross-section of a single sheet of PVF_2 film. Nickel metallization provides electrical contact for the material. Zig-zag chain of molecules is responsible for the material's properties. Polarity is shown for one stretch direction. Note the polarity of the electric field.

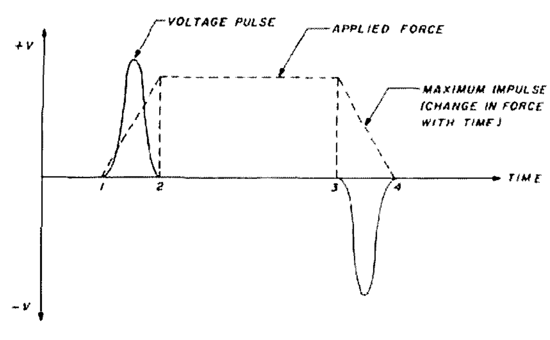


fig. 2. Rate effect of stretched PVF_2 film. Voltage pulses are maximum for maximum rate of change in applied force. No voltage output exists for a steady-force condition.

stant and no output voltage is produced since charge redistribution has occurred. The declining impulse slope at the right hand side of the figure (between points 3 and 4) produces a corresponding negative-going voltage pulse.

Since voltage pulses, rather than a static DC voltage, are produced when PVF_2 film is stretched, the voltage pulses must be observed on an oscilloscope. Using one, I observed peaks as high as ±10 volts when a sample of 52-μm thick material was given a sharp tug. The output impedance of the material is quite high (over 10 megohms) since the actual current flow in the material is extremely small.

PVF_2 sheets are available in various thicknesses. Certain thicknesses are better suited for specific applications: for *mechanical-to-electrical* conversions, *thick*, uniaxially stretched material works best because the greater the material's thickness, the greater the voltage output. For *electrical-to-mechanical* conversions, *thin*, uniaxially stretched material is preferred because the thinner the material, the greater the mechanical displacement.

PVF_2 is easy to work with

To cut PVF_2 film to the required dimensions, begin by taping the edges of the sample flat. Using a metal straightedge as a guide, cut the film with an X-ACTO™ knife, using a steady motion. *Don't use scissors*; they tend to deform the cut edges.

Electrical connection to the material can be made in several ways. The method I used, certainly one of the most economical, is to use a copper foil tape with conductive adhesive. I used a 3M® tape, No. 1181, which sells for approximately $12.00 for a roll 18-yards

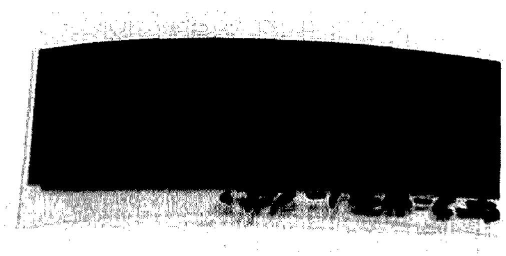


fig. 3. A sample of 110 μm PVF_2 film. The serial number indicates the film type. Note the + sign and the right arrow indicating the instantaneous polarity and stretch direction of the top metallized surface.

(16.2 meters) long and 3/8-inch (0.94 cm) wide. Because very little foil is required, a single roll can be used for many applications and should last for quite a while.

I found that No. 30 wire-wrap wire was the easiest to use because it is thin enough not to interfere with any mechanical *actions* and is very flexible. When soldering, don't attempt to solder directly to the PVF_2 film; this will destroy the material and could liberate toxic hydrogen fluoride gas. Instead, first solder the wire to the copper foil tape and then attach the foil tape to the material. Generally it's best to tin the foil tape first and then bring the wire to the tape. Once the lead is connected to the foil tape, remove the protective backing and apply the foil tape to the material, using finger pressure. While the tape's adhesive is quite strong when the tape is pulled along its length, it can be carefully peeled off if necessary.

Keep the wire leads at a 90-degree angle to the material's "stretch direction." In some instances I found myself tugging on the wires and generating a voltage when I was not expecting it. *The right angle*

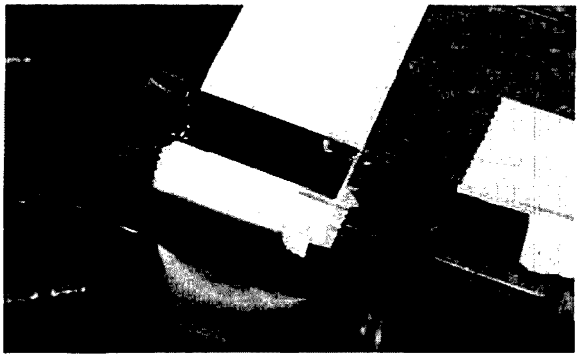


fig. 4. View of the keyer paddle arrangement. Note that the wire leads are oriented 90 degrees to the material's stretch axis. The facial tissue acts as a shock absorber. The two pieces of drafting tape are visible at the right edge of the material.

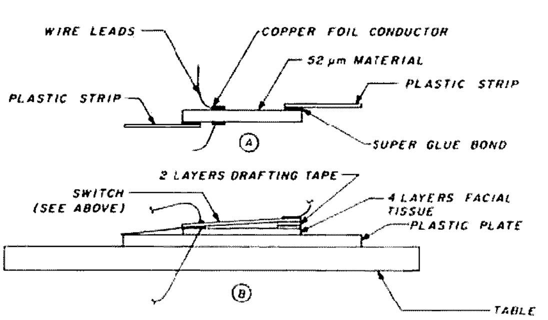


fig. 5. Side view of keyer paddle arrangement. (A) shows switch (note that plastic strips are glued at top and bottom for maximum elongation); (B) shows sidetails of entire set-up. The facial tissue acts as a shock absorber; drafting tape elevates one end of film.

approach prevents force from being applied along the stretch axis.

Figure 3 illustrates a typical sample of the material. Note the serial number indicating the film's thickness (110 μm in this case) and the markings on the lower right hand corner. The + sign indicates the positive polarity side of the material; the stretch direction is indicated by the arrow at the right. When cutting, be sure to keep track of polarities and stretch directions at all times. I found it best to use drafting tape (such as Scotch® brand No. 230) to label and hold down all cut pieces because the tape adheres well but can be easily removed without wrinkling the film. Masking tape is too sticky and too difficult to work with, especially for thin samples of material.

Morse code keyer paddle

After experimenting with PVF_2 film for a few months I decided it was time to try to produce something useful. Intrigued by a sketch that showed the positive- and negative-going pulses of a piece of stretched film when it is struck, I immediately thought of using the film in a Morse code key or keyer paddle. With the material, a lightweight, nonmechanical device could be constructed to fit any size constraint. Its performance could be varied through electronic, rather than mechanical adjustments. Because I didn't have enough spare film to construct the keyer paddle I had first envisioned, I decided instead to build a workable circuit and mechanical configuration that could be easily modified and extended by any Amateur interested in pursuing this topic.

Because thicker PVF_2 films are better suited for mechanical-to-electrical conversions than thinner films, I decided to try working with the 52 μm piece supplied with the sample package. **Figure 4** shows the experimental keyer paddle (actually one side of a keyer paddle arrangement) and **fig. 5** shows construction details.

A strip of 52 μm material 1-1/8 inch (2.81 cm) long and 1/2 inch (1.25 cm) wide was glued with Super Glue® (an Eastman 910® type of adhesive) to two plastic strips cut from an old blister pack. I found that only a small amount of Super Glue is needed to bond the plastic to the material and that air is necessary for proper drying. Facial tissue is used as a shock absorber to prevent damage to the underside of the film. The two layers of drafting tape raise one end of the material to allow room for movement. The plastic plate serves merely to protect the table from the vise and C-clamp used to secure the prepared 52 μm strip. The wire-wrap wires are then routed to the paddle's electronic circuitry.

Because the keyer paddle produces only positive and negative pulses as it is depressed and released, the electronic circuit shown in **fig. 6** is needed to produce the steady key-up and key-down conditions necessary for Morse code transmission. The circuit amplifies and buffers the input transitions, detects the polarity of the transitions, and sets or resets a flip-flop at the beginning and end of a keying interval. A 555 timer serves as an audio oscillator to monitor the keying.

U1 is a LF353 BIFET op amp characterized as having an extremely high input impedance and an extremely low input offset current. Because PVF_2 film is unable to produce even moderate amounts of current, it is necessary to use the BIFET-type op amps with their low input offset currents. At first I tried using a conventional 741 op amp and found that both the input impedance and input offset current were incompatible with the material. R1 and R2 set the gain of U1A to approximately −12.2, and U1B uses R3 and R4 to set its gain at approximately +121. Resistors R5 through R9 adjust the op amp's offset voltage. Ad-

just R7 and R8 for zero volts at the junction of steering diodes CR1 and CR2 with no input applied.

U2 is an LM339 voltage comparator. U2A acts as a detector for positive transitions and U2B as a detector for negative transitions. The respective voltage thresholds are set by resistors R12 and R15. By trial and error I found that setting the voltages at pins 4 and 11 of U2 to + and −1.5V seemed to work best for keyer applications. In general, I found that the threshold voltages should not be set below 0.6V because if they are, the circuit will become susceptible to false triggering, either through keyer paddle vibrations or through other electrically-induced signals. Because the thresholds are completely variable, it's possible to vary the paddle's "weighting" or how hard the film must be struck to produce the required voltage pulses. Compare this to conventional mechanical keyer paddles where the "feel" is almost impossible to vary and the advantages of using PVF_2 film are immediately obvious.

U3 is a LM1489 EIA-to-TTL level converter commonly used for RS-232 applications. It serves here as a bipolar-to-unipolar converter for operating the logic circuits of U4 and U5. Because U3's output is at a TTL level, R16 and R17 are needed to pull up the output voltage during a logic "1" condition.

U4A and U4B are configured as inverters to compensate for U3's logic inversion. U5A acts as an SR flip-flop, set by positive-going transitions at the material's output leads and reset by the negative-going transitions once the material is released. Because the transition polarities are dependent upon the direction the material is stretched, it may be necessary to either reverse the input leads or turn the piece of material over in order to find the correct triggering polarity. (If your keying pattern sounds "inverted," you'll know that the wires need to be switched.)

U6 acts as an astable multivibrator when the top of R19 is at 5V due to U5A pin 1 being held at a logic "1." R19, R20, and C5 form the IC's timing circuit and determine the tone's frequency. R18 acts as a pull-up and load resistor, and C6 acts as a coupling capacitor. If you're going to be driving other circuits, U6 is not needed.

When working with the BIFET op amp, keep all component leads as short as possible. Keep the connecting leads from the keyer paddle shielded and as short as possible. Given the high input impedance of the BIFET op amps, it is very easy to turn the keyer paddle into an antenna if you're not careful. Here again, the variable comparator thresholds do give you some freedom in preventing false triggering. Use whatever bypass capacitors are required for your particular application. Remember, keep the op amp bypassed!

Once the circuit was breadboarded, the keyer paddle leads were connected. When I first touched the material's surface I started to pick up a 60 Hz hum, indicating that some type of isolation would be necessary. Since the keyer paddle was partially obscured by the vise and C-clamp I was forced to use a cotton swab as a cushioned striker for the material. Its cotton tip protected the films' surface from scratches and puncture, but was sufficiently pointed to produce an adequate impulse for keying purposes. Although holding the cotton swab was slightly awkward, I was able to realize reliable keying at speeds of 10 to 15 WPM. In a normal keyer paddle configuration a "clean" layout should allow you to operate the paddle as you would any other keyer. In an actual keyer paddle it would be best to use a piece of protective plastic over the paddle's surface to prevent scratching the material and to place a "dimpled" or rounded object between the plastic sheet and the keyer paddle to simulate the impulse actions of the cotton swab.

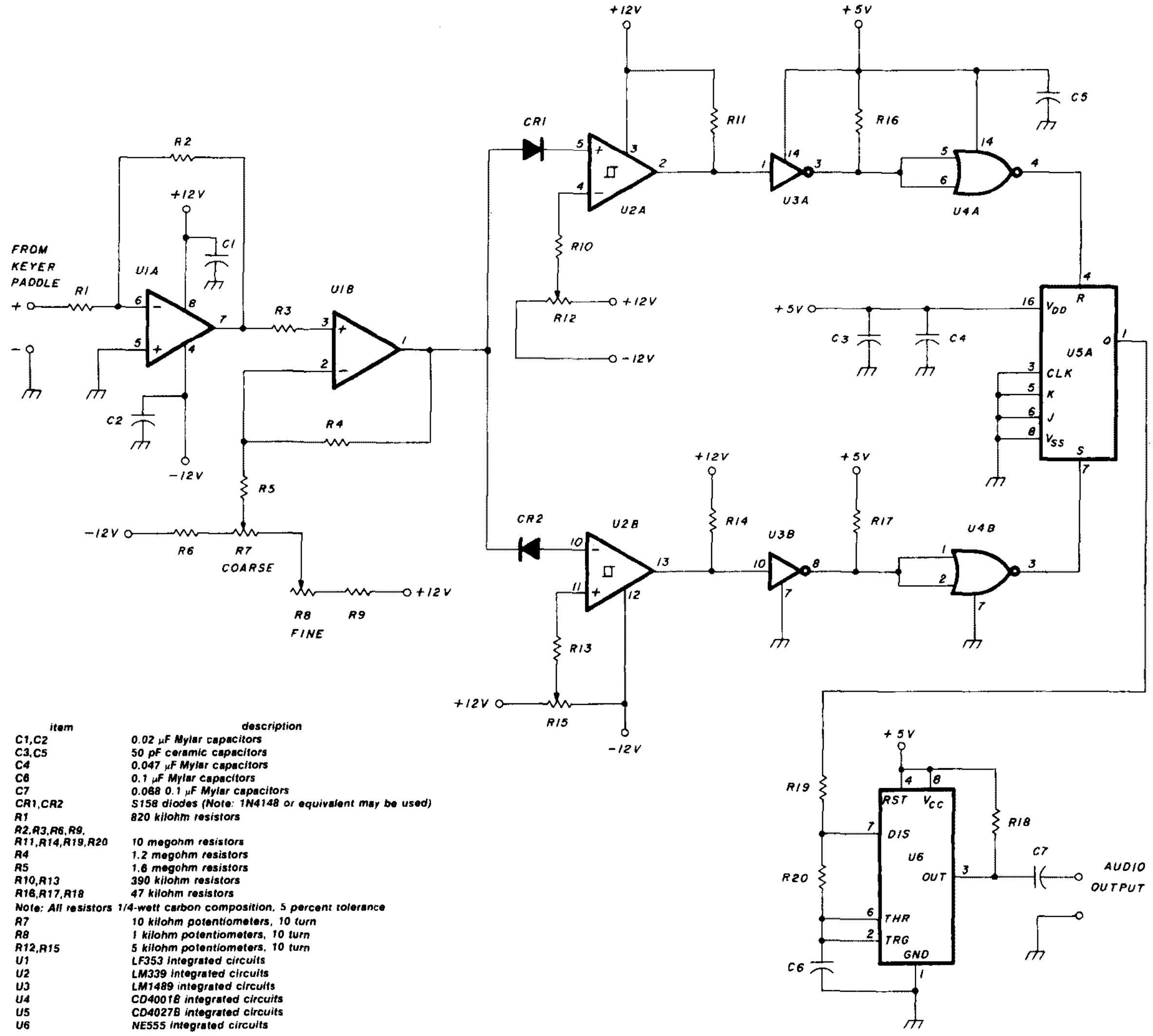
I did try using a sample of 110 μm material for the keyer paddle but found that, in general, it was too sensitive for keyer paddle applications. In fact, the 52 μm sample used in the keyer paddle was so sensitive (with 0.6V thresholds) that I could knock on the table and trigger the circuit to the "on" state. I'm sure you'll find 52 μm material to be more than sufficiently sensitive for any keyer paddle applications you may want to investigate.

film works as heat sensor, too

Although mechanical-to-electrical and electrical-to-mechanical conversions have been my primary topic for research, it is also possible to use PVF_2 film to sense infrared (heat) radiation. In this case the material produces a voltage when a changing heat source is brought near the material. Waving a match in front of a sheet of PVF_2 will cause the film to produce a voltage. Should you decide to experiment with PVF_2 film, it is important to remember not to burn the PVF_2 film or expose it to temperatures in excess of 572 degrees F (300 degrees C) because toxic hydrogen fluoride gas will be liberated.

further applications for ham radio

I hope by now I've gotten your mental motor running and you're thinking about other possible applications for PVF_2 film. As you ponder, keep in mind the cost of the material and the type of mechanical-electrical conversion you want to perform. Because the cost of PVF_2 film ranges from 17.5 cents/cm^2 for 9 μm material to 75 cents/cm^2 for 110 μm material, the average ham might find large sheets of this material too expensive for use in any one application.



item	description
C1,C2	0.02 μF Mylar capacitors
C3,C5	50 pF ceramic capacitors
C4	0.047 μF Mylar capacitors
C6	0.1 μF Mylar capacitors
C7	0.068 0.1 μF Mylar capacitors
CR1,CR2	S158 diodes (Note: 1N4148 or equivalent may be used)
R1	820 kilohm resistors
R2,R3,R6,R9, R11,R14,R19,R20	10 megohm resistors
R4	1.2 megohm resistors
R5	1.6 megohm resistors
R10,R13	390 kilohm resistors
R16,R17,R18	47 kilohm resistors
Note: All resistors 1/4-wett carbon composition, 5 percent tolerance	
R7	10 kilohm potentiometers, 10 turn
R8	1 kilohm potentiometers, 10 turn
R12,R15	5 kilohm potentiometers, 10 turn
U1	LF353 integrated circuits
U2	LM339 integrated circuits
U3	LM1489 integrated circuits
U4	CD4001B integrated circuits
U5	CD4027B integrated circuits
U6	NE555 integrated circuits

fig. 6. Schematic diagram of the keyer paddle circuit. The paddle's output signal is amplified by U1, threshold-compared by U2, converted to unipolar logic levels by U3, inverted by U4, and held in a key-up or key-down condition by U5. U6 acts as an audio oscillator.

Take heart; most applications require only a small amount of material, so one purchase should last quite some time. For larger purchases, sharing a purchase with a friend or with a club might make acquisition more affordable. The 52 μm keyer paddle described in this article required a piece measuring only 0.56 inch^2 (3.5 cm^2), at a cost of $1.23. Because the material is fairly new, the cost per cm^2 should decrease steadily with time.

For hams who have projects in which space and weight are important (such as QRP rigs and hand-held radios) the advantages of using PVF_2 film in microphones or speakers may more than offset the cost of the film. In fact, for QRP applications, for example, where CW is the primary medium of communications, a PVF_2 film speaker could be built inexpensively. This speaker would be physically small and lightweight and would consume very little power. It could be custom-tailored to operate over any desired band of audio frequencies. Other applications such as "do-it-yourself" keypads for equipment are also possible.

PVF_2 film has been used in transducer applications at frequencies up to 9.36 GHz. One can easily envision different types of antennas and RF pick-up schemes that Amateurs could easily build. The literature does make reference to vidicons or video "tubes" for cameras using PVF_2 film pieces to form an imaging

area so SSTV and ATV enthusiasts may wish to read more about this. Believe me, I could easily ramble on about potential applications for pages on end, but my advice to you is read all of the applications literature available and try your own experiments.

Applications literature for PVF_2 film is available from KYNAR Piezo Group, Pennwalt Corporation, 900 First Avenue, P.O. Box C, King of Prussia, Pennsylvania 19406-0018.

acknowledgements

Special thanks to J. Victor Chatigny at Pennwalt for his comments and assistance and to my friend Bill Drake, for his help with the photos.

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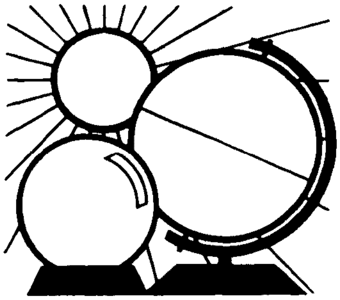
ham radio



DX FORECASTER

Garth Stonehocker, KØRYW

absorption affects signal level

Last month's column included a discussion of the effect of absorption on signal levels. Under *normal* conditions a transmitted wave leaving the antenna at an effective radiated power level of 1000 watts, traversing the atmosphere twice and the ionosphere once, might arrive at the receive location over 170 dB down (weaker) and still be copyable. During the winter a lower absorption level and a quieter (or undisturbed) geomagnetic field results in enhanced conditions for the lower frequency bands. Occasionally, however, a condition called the *winter absorption anomaly* appears to reverse the trend.

This anomaly causes higher than usual attenuation of signals crossing a 500 to 2500 km wide strip between 60 and 70 degrees North latitude (or more precisely, 40 to 55 degrees *geomagnetic)*. The region is known as the mid-latitude trough, and the effect lasts for several days at a time. The trough is an actual depression in electron density of the F region and a higher than usual level of D and E region ionization. It is the latter condition specifically that accounts for greater wintertime absorption levels on the lower bands. One can expect it to occur from November through February and to be more severe during times of greater geomagnetic activity. However, there is no *short* term (i.e., hour or day) relationship to solar flux or geomagnetic activity indices. The extra ionization is related to electron particles entering the auroral zone. No applicable data collection or warning system is currently available for forecasting the anomalous absorption.*

last-minute forecast

The 27-day solar (activity) cycle variations which have increased over those of the summer and fall, are expected to be even greater in December. Look for a maximum in activity to occur on approximately the sixth of the month, with a corresponding minimum on the twentieth. A solar flux of 80 represents an average value for this part of the solar cycle. The effect of this increased solar activity means that the 10 through 30 meter bands will be useful for communicating between North America and South Africa, South America and Australia over trans-equatorial paths during the first two weeks in December. The latter weeks will favor the lower frequency bands — 30 through 160 meters. Tune in to WWV at exactly 18 minutes after the hour to receive their daily broadcasts of solar activity and geomagnetic field information useful in determining HF/VHF band conditions. Also during this month, both atmospheric noise and ionospheric absorption will be down, resulting in useful *daytime* DX conditions for the 40 and 80 meter bands.

The Geminid meteor shower, which reaches its peak on December 13-14, provides the richest and most reliable display of the year, with rates of 60-70 per hour (measured mainly by radio because of the poor weather in December). Also, a smaller portion of the shower (15-20 per hour) is observed on December 22. Lunar perigee and full moon occur on December 7th and 8th, respectively. Winter solstice — the longest night — occurs on December 21.

*Reports of this absorption would be useful to me. Please write to Garth Stonehocker, Route 2, Box 45, Earlysville, Virginia 22936.

band-by-band summary

Ten, fifteen, and twenty meters will be open from morning to early evening almost every day, and to most areas of the world. The openings on the higher of the bands will be shorter and will occur closer to local noon. Trans-equatorial propagation on these bands will be more likely towards evening during conditions of high solar flux and a disturbed geomagnetic field.

Thirty and forty meters will be useful almost 24 hours a day. Daytime conditions will resemble those on 20 meters but shorter skip and signal strength may decrease during midday on some days — days coinciding with high solar flux values or anomalous absorption. Nighttime use will be good except after days of very high MUF conditions. Generally the usable distance is expected to be somewhat greater than that achieved on 80 at night.

Eighty and one-sixty meters are the nighttime DXer's bands, especially this time of year. The bands open just before sunset and last until the sun comes up on the path of interest. Except for daytime short-skip signal strengths, high solar flux values don't affect these bands much. Some days the anomaly will affect day and night signal strength.

DECEMBER

WESTERN USA

GMT	PST	N (ASIA FAR EAST)	NE (EUROPE)	E (S. AFRICA)	SE (S. AMERICA)	S (ANTARCTICA)	SW (NEW ZEALAND)	W (OCEANIA AUSTRALIA)	NW (JAPAN)
0000	4:00	30	30	15	15	15	10	15	20
0100	5:00	20	30	15	10	15	10	10	20
0200	6:00	20	30	20	10	15	15	10	20
0300	7:00	20	30	20	10	15	15	15	20
0400	8:00	20	30	20	10	15	15	15	20
0500	9:00	20	_40_	20	15	15	15	15	20
0600	10:00	_30_	_40_	20	15	20	_20_	15	_30_
0700	11:00	_30_	_40_	_20_	15	20	_20_	_20_	_30_
0800	12:00	_30_	_40_	_20_	_15_	_20_	_20_	_20_	_30_
0900	1:00	_30_	_40_	_20_	_20_	_20_	_20_	_20_	_30_
1000	2:00	_30_	_40_	_30_	_20_	_20_	_20_	_20_	_30_
1100	3:00	_30_	_40_	_30_	_20_	_20_	_20_	_20_	_30_
1200	4:00	_30_	_40_	_20_	_20_	_20_	_20_	_20_	_40_*
1300	5:00	_40_	_40_	_20_	_20_	_20_	_20_	_20_	_40_
1400	6:00	_40_	30	_20_	_20_	_20_	_20_	_20_	_40_
1500	7:00	_40_	30	15	_20_	_20_	_20_	_20_	_40_
1600	8:00	_40_	30	15	_20_	_20_	15	_20_	_40_
1700	9:00	_40_	30	15	_20_	15	15	_20_	_40_
1800	10:00	_40_	30	15	15	15	15	15	30
1900	11:00	_40_	30	15	15	15	15	15	30
2000	12:00	_40_	30	15	15	15	15	15	20
2100	1:00	_40_*	30	15	15	15	15	15	20
2200	2:00	30	30	15	15	15	15*	15	20
2300	3:00	30	30	15	15	15	10	15	20

MID USA

GMT	MST	CST	N (ASIA FAR EAST)	NE (EUROPE)	E (S. AFRICA)	SE (S. AMERICA)	S (ANTARCTICA)	SW (NEW ZEALAND)	W (OCEANIA AUSTRALIA)	NW (JAPAN)
0000	5:00	6:00	30	30	15	15	15	15	15	20
0100	6:00	7:00	40	30	15	20	15	15	15	20
0200	7:00	8:00	40	30	15	20	15	15	15	20
0300	8:00	9:00	40	30	20	20	15	15	15	20
0400	9:00	10:00	40	30	20	20	15	20	15	20
0500	10:00	11:00	_40_	_40_	20	20	15	20	20	30
0600	11:00	12:00	_40_	_40_	_20_	20	20	20	20	30
0700	12:00	1:00	_30_	_40_	_20_	_20_	_20_	20	20	_30_
0800	1:00	2:00	_30_	_40_	_20_	_20_	_20_	_20_	20	_30_
0900	2:00	3:00	_30_	_40_	_20_	_20_	_20_	_20_	_20_	_30_
1000	3:00	4:00	_30_	_40_	_20_	_20_	_20_	_20_	_20_	_30_
1100	4:00	5:00	_30_	_40_	_20_	_20_	_20_	_30_	_20_	_30_
1200	5:00	6:00	20	_40_	_20_	_20_	_20_	_30_	_20_	_30_
1300	6:00	7:00	20	30	_20_	_20_	_20_	_20_	_20_	_40_
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1500	8:00	9:00	20	30	15	15	_20_	_20_	_20_	_40_
1600	9:00	10:00	20	30	15	10	_20_	15	_20_	_40_
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1800	11:00	12:00	30	30	15	10	15	15	15	30
1900	12:00	1:00	30	30	15	10	15	15	15	20
2000	1:00	2:00	30	30	15	10	15	15	15	20
2100	2:00	3:00	30	30	15	15	15	15	15	20
2200	3:00	4:00	30	30	15	15	15	15	15	20
2300	4:00	5:00	30	30	15	15	15	15	15	20

EASTERN USA

GMT	EST	N (ASIA FAR EAST)	NE (EUROPE)	E (S. AFRICA)	SE (CARIBBEAN S. AMERICA)	S (ANTARCTICA)	SW (NEW ZEALAND)	W (OCEANIA AUSTRALIA)	NW (JAPAN)
0000	7:00	40*	30	15	15	15	15	15	20
0100	8:00	40	30	15	20	15	15	15	20
0200	9:00	40	30	20	20	15	15	15	20
0300	10:00	_40_	30	20	20	15	15	15	30
0400	11:00	_40_	30	20	20	20	20	15	30
0500	12:00	_40_	_40_	20	20	20	20	20	_30_
0600	1:00	_40_	_40_	_20_	20	20	20	20	_30_
0700	2:00	_40_	_40_	_20_	_20_	_20_	20	20	_30_
0800	3:00	_40_	_40_	_20_	_20_	_20_	_20_	20	_30_
0900	4:00	_30_	_40_	_20_	_20_	_20_	_20_	_20_	_30_
1000	5:00	_30_	_40_	_30_*	_20_	_20_	_20_	_20_	_40_
1100	6:00	_30_	_40_	_20_	_20_*	_20_	_20_	_20_	_40_
1200	7:00	_30_	_40_	_20_	_15_	_20_	_30_*	_20_	_40_
1300	8:00	20	_30_	15	_15_	_20_	_20_	_20_	_40_
1400	9:00	20	_30_	15	15	_20_	_20_	_20_	_40_
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1600	11:00	20	30	15	15	_20_	15	_20_	_40_
1700	12:00	20	20	15	10	15	15	15	_40_
1800	1:00	30	20	15	10	15	15	15	30
1900	2:00	30	20	15	10	15	15	15	30
2000	3:00	30	20	15	10	15	15	15	30
2100	4:00	30	20	15	10	15	15	15	20
2200	5:00	30	20	15	15	15	15	15	20
2300	6:00	30	30	15	15	15	15	15	20

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during "normal" hours.

*Look at next **higher** band for possible openings.

ham radio TECHNIQUES

Bill Orr W6SAI

huff-duff revisited

"Huff-duff" is a World War II term for *high-frequency direction finding*, a technique vital for locating enemy transmitting sites. A considerable amount of money and energy were expended in maintaining this essential service.

HF-DF (the modern term) was developed about 1938, principally by the Australian Ministry of Civil Aviation. Many direction finding stations were installed on that continent. But as aeronautical radio communication in the United States gradually shifted over to VHF bands, the extensive use of high-frequency direction finding became unnecessary. America's entry into World War II, however, changed this, and accelerated development of new equipment was quickly undertaken by the FCC and the United States Army Air Corps. After the war, some of this equipment showed up in the surplus markets as the SCR-291, which was largely ignored by Radio Amateurs.

Amateur direction finding today

Interest in HF-DF is slowly growing among Radio Amateurs. Those who experiment with this interesting hobby soon learn that a new set of ground rules applies, especially with regard to long distance ionosphere-reflected signals.

Direction finding involves the determination of a *line of direction* between a receiver and a distant transmitter. The intersection of two lines of direction from widely separated receivers establishes the location of the transmitter. This technique works well on ground wave signals, but "skip" signals reflected from the ionosphere undergo alterations that greatly increase the difficulty of determining a line of direction. As long as ground wave signals are used for direction finding, accurate bearing readings are relatively easy to obtain. DF work has been accomplished with good results on VHF (the 2-meter band, for example) and on ground wave signals on the 160-meter band. But Amateurs have had little experience with DF work on the DX bands with long-distance signals.

DF principles

The loop antenna is the device that forms the basis of all low- frequency DF work, such as demonstrated by the radio ranges operating in the 200 kHz to 500 kHz portion of the spectrum. Nondirectional (transmitting) ground stations are used as a signal source on which bearings are taken by a special receiving antenna. In most instances, a simple loop antenna is used (**fig. 1**). If the dimensions of the loop are small compared to wavelength, the pickup pattern of the loop is as shown in **fig. 2**. Maximum pickup occurs when the radio wave travels parallel to the plane of the loop, cutting first one side then the other. If the loop is at right angles to the direction of wave travel, the radio wave cuts both sides of the loop simultaneously and the resultant output voltage of the loop is zero. The resulting pickup response of the loop is thus a figure-eight pattern and there are two loop positions for which the output is zero and two for which it is a maximum. Note that the loop pattern is opposite to that of a Quad loop, or other large loop, wherein the plane of maximum signal response is at right angles to the plane of the loop.

Note also that the nose of the loop pattern is extremely broad and the pattern nulls are very sharp. It is the pattern nulls which are useful for direction-finding work.

Operation of the loop direction finder is simple. The receiver is tuned to the desired station and the loop is rotated until the signal disappears. The line of direction is then compared with either magnetic or true north, and the results plotted on a map. Two such bearings, from separated locations, constitute a *fix*, or the location of the signal source.

the HF-DF problem

The most important alteration to the radio wave as it is reflected from the ionosphere is a change in polarization. The electric field of a radio wave exerts a force on electrons in the ionosphere, in addition to the force exerted on them by the magnetic field of the earth. The resulting action of the two forces causes an elliptical electron motion that causes the polarization of the exciting wave to be elliptical. Measurements made in 1931 showed that regardless of the initial polarization of the transmitted wave, the reflected wave polarization was elliptical for medium distances, with the possibility that polarization could be elliptical or horizontal in the case of long- distance waves striking the earth at a low angle of incidence.

In addition to polarization changes, the reflected radio wave exhibits fading, probably as a result of two or

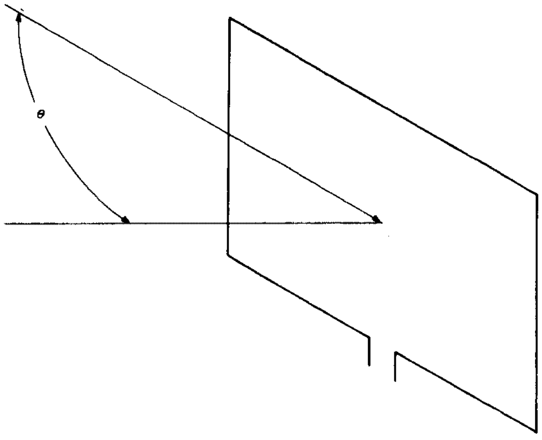

fig. 1. Output voltage of small loop is proportional to the cosine of the angle θ between plane of the loop and signal vector.

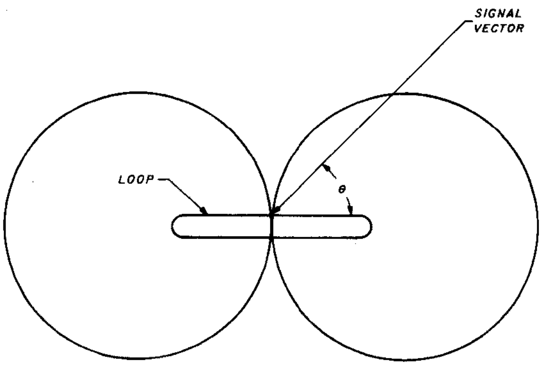

fig. 2. Pickup pattern of small loop. Note nulls at right angles to plane of loop. Maximum pickup occurs when signal vector is parallel to plane of loop.

more reflection paths having slightly different lengths. The different paths come about because of irregularities in the refracting layers of the ionosphere. More often than not the paths are sufficiently separated so that the received signal seems to be coming from two separate directions.

deviations from the great-circle path

It is widely assumed that long-distance signals follow a great-circle path from transmitter to receiver. Amateurs frequently place considerable faith in great-circle maps when aiming their directive antennas at an exotic DX station. Experienced operators, however, have long known that this confidence is not always justified. With a small beam antenna having a "fat" forward lobe, azimuthal deviation of the DX signal is of little importance. Users of highly directive beams, on the other hand, have found that deviations of 10 to 20 degrees from the great-circle path of DX signals are common over short periods of time, and deviations as much as 90 degrees have been noted for signals appearing from the antipodal area of the world.

In the case of sporadic-E layer reflected signals, reflection can occur from many "clouds" of reflective regions and multiple echoes are present on a signal propagated in this fashion. Experimenters have noted that bearing deviations up to 20 degrees are present on echo signals.

the loop antenna for DF work

"Top Band" DXers have used small loop antennas for reception for many years. Generally speaking, the loop is not used to DF the received signal, but to null out local interference (such as power line noise, or TV oscillators) to achieve improved signal-to-noise ratio. *However, most users of loop antennas have found out that they are useless as direction finders for ionospherically reflected signals.*

I have used a simple loop antenna for tracking thunderstorm interference on 160 meters, but the directional nulls were broad and pointing only to a very general area of thunderstorm activity.[1] No success was had DFing long-distance signals on this band.

the Adcock DF antenna

In 1919 a patent was granted to F. Adcock, an Englishman who developed an HF direction finding antenna that was not susceptible to polarization error of the reflected radio wave. The antenna was sensitive only to the vertically polarized component of the wave (**fig. 3**). Amateurs will recognize this device as similar to, though smaller than, the famous W8JK beam antenna. If the spacing between antenna elements is small compared to the wavelength, the voltage output of an Adcock antenna is a function only of the height above ground and the angle between the vertical plane containing the elements and the plane of polarization of the radio wave. Until World War II, most practical direction finders used various versions of the Adcock antenna. The old SCR-291 DF equipment used a version of the Adcock antenna, plus a "sense" antenna that removed the 180-degree ambiguity caused by the Adcock array. An oscilloscope was used to view the SCR-291 output, which provided a DF signal picture as illustrated in **fig. 4.**

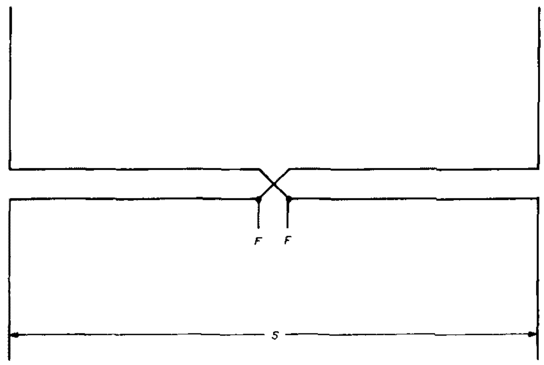

fig. 3. Balanced Adcock antenna system which responds to vertically polarized waves. Spacing S is small in terms of wavelength.

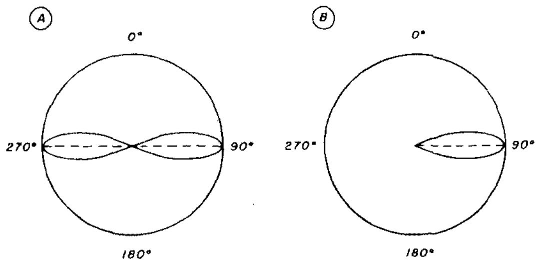

fig. 4. (A) Elongated figure-eight pattern provides line of direction in SCR-291 direction finder. (B) Sense antenna changes "leaf" pattern to cardioid to resolve signal ambiguity.

For accurate DFing, up to 100 bearings were taken at various times during a 24-hour period, with particular attention to bearings taken at sunrise and sunset. Bearings from two or more widely separated DF receivers were taken, smoothed, and then plotted on a map, thereby indicating an area in which the transmitter was located with a reasonable degree of probability.

The Adcock system has been supplanted by superior devices, but these are large and beyond the cost of the average Amateur. But when properly built and used, the Adcock antenna is a valuable addition to the Amateur interested in HF direction finding.

building an Adcock array

I've seen only one article dealing with the construction of an Adcock antenna.[2] In brief, the Adcock array consists of two vertical antennas, closely spaced with respect to wavelength and connected out-of-phase, as shown in fig. 5. Each antenna can be shorter than its physically resonant length, with system resonance established by placing a simple tuner at the operating position. Using a 0.1 wavelength design for a 20-meter Adcock array provides element spacing of about 7 feet 6 inches. Signal pickup depends upon element length. In order to make a compact array, 0.1 wavelength elements were chosen.

The array elements are cross-connected with a short length of open-wire TV "ladder line." The feedpoint is taken at the center of the line. A second length of ladder line leads from the array down to the antenna tuner at the operating position. Care should be taken to keep the line clear of nearby metallic objects.

The tuner consists of a resonant circuit with an inductive pickup link coupled to the receiver via a coaxial line. Balancing capacitors are connected across the tuning capacitor to enhance the signal null. It's best to locate the tuner directly below the antenna, if possible, to ensure a good signal null.

The first model antenna was built on a wood frame which proved to be rather unsteady. A new model, in which aluminum tubing is used for the framework supporting the antennas, is under way.

Antenna system resonance can be established with a dip meter or by adjustment of the tuner on a local signal. When resonance is found, the antenna is rotated for the best null and the balancing capacitors (C2,C3) are adjusted to enhance the null. The antenna exhibits a figure-eight pattern, with the nulls at right angles to the plane of the antennas.

It may be difficult to find an exact null when trying to determine a line of direction to a signal because of background noise obscuring the null. In this case, the null can be estimated by noting the position of the antenna at an equal signal position on each side of the null. The angular difference between the readings is divided in half to find the actual null.

In my case, the experimental Adcock array is turned by a small TV rotator because the wind resistance of the antenna is quite low. The array and read-out of the rotator are adjusted so that "North" on the indicator corresponds to true North (not magnetic North).

using the Adcock array

It's best to practice on local signals and one-skip signals. (I use stations in the Denver area for DF practice.) Beware of local reflections from hills and from guy wires, gutters, and other nearby metallic objects. Use a map to determine the true direction of your practice stations, and make multiple

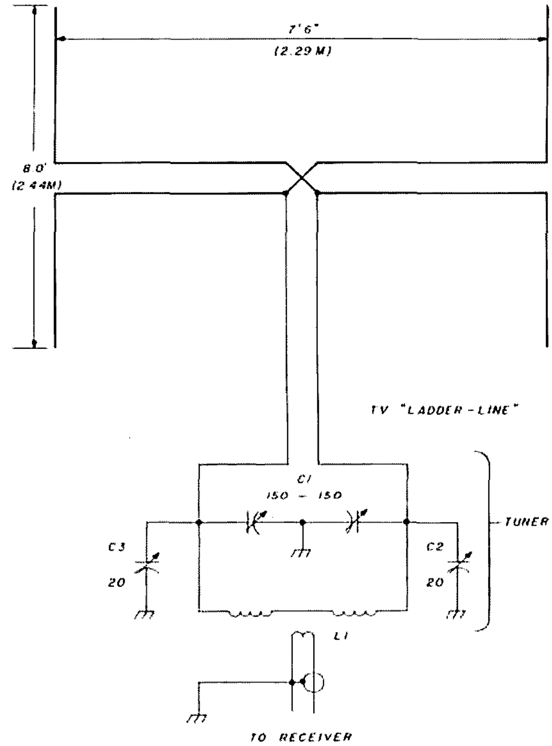

fig. 5. Simple Adcock array for 14-21 MHz operation. C1-L1 resonates system to listening frequency and C2, C3 are used to enhance signal null. (Patterned after design by W1YNC, *QST*, August, 1975, page 30.)

readings to obtain meaningful results. Remember, the loop provides an ambiguous reading with an error factor of as much as 180 degrees if you don't keep your wits about you. Pretty soon you can "spot" a station with reasonable directive accuracy before you hear its location or look it up in the Callbook.

If you have a general coverage receiver, you can venture out of the ham bands. The 15-MHz broadcast band is a good one on which to try DF experiments. A prime choice is WWV in Colorado, as are the Voice of America outlets, the BBC, HCJB in Quito, Ecuador, and Radio Japan. DFing requires a certain amount of expertise if your efforts are to produce meaningful results, but with practice you'll become more skillful. Then when 6L6GX opens up from Lower Bulltravia, you can tell your DX friends, "Don't waste time on him. His heading doesn't check with where he says he is!" (This reminds me of the story of an infamous DXer who appeared to have worked the world from many exotic locations until his adventures were finally branded as fraudulent. DXCC credit was revoked for the thousands of "new country" contacts he had provided for gleeful DXers. After the excitement had died down, he resurfaced one day, working portable from New York City. While in QSO with someone, a hard-bitten DXer zero-beat the fraud and very slowly sent: "Gee! Do you really think he's there?!")

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ham radio

quasi-bilateral IF for transceivers

Try this old technique to save space in new circuits

Oldtimers will remember the bilateral IF circuits that were popular some years ago. The intent of their designers was to avoid duplication of circuitry in the receiver and transmitter stages in compact transceivers. Although a few manufacturers, notably Side-Band Engineers, made use of bilateral IF stages, the technique never fully caught on with homebrewers, possibly because of the overall complexity and parts count involved with the early bipolar circuits, and the tendency towards instability exhibited by some designs. Orr's *Radio Handbook* carried a brief description of a bilateral IF arrangement similar to the SBE circuit.[1] The bilateral IF has gone out of vogue in recent years with the advent of modern IC technology, high density PC board layouts, and inexpensive semiconductors.

When I decided to build a compact 160-meter transceiver the use of such a scheme seemed ideal to conserve space. Unfortunately, all of the designs appearing in my various handbooks left much to be desired and were in fact quite dated. The final circuit that was used successfully in my project is presented here.

My objectives for a simple, effective circuit were met through the use of IC and passive doubly balanced diode ring mixer components available at low cost and small in physical size. While my circuit is not a true bilateral design, it represents a unique approach to a shared receive/transmit IF that I haven't seen used. (Benjamin Vester, W3TLN, was on this track when he described his pocket-sized solid-state 20-meter transceiver in the ARRL's *Sideband Handbook*.[2] I had chosen an IF frequency of 455 kHz to make use of some old Collins mechanical filters I found in my junkbox. Crystal filters of more recent vintage should work as well or better because of high insertion loss associated with the early mechanical filters I used. My 160-meter transceiver is single conversion, and a low IF frequency was not a concern.

The first stage of the IF, a Mini-Circuits Lab DBM, does double duty as the balanced modulator for transmit and as the receiver first mixer. During receive, the preselector output (1.8 MHz) is coupled to the IF port of the DBM while the LO port is driven by the VFO, with the 455 kHz IF resulting. During transmit the IF port is driven with amplified microphone audio for SSB operation. The LO port is driven by the BFO. Note for CW generation during transmit a DC bias on the mixer IF port allows BFO energy to feed through the mixer; of course the BFO frequency must be shifted to within the passband of the filter for proper transmitter operation. SSB audio is coupled to the IF port through an RF choke so as not to affect receive operation, while the value of the coupling capacitor from the IF port to the receiver preselector has insufficient reactance to adversely load the microphone audio.

This confirguration admittedly compromises the dynamic range of the DBM. Termination impedance of the IF port of diode ring mixers is critical and directly affects the insertion losses and dynamic range. Care was taken to provide a diplexer at the IF output (the manufacturer's designated RF port for the DBM), and with the use of a 3-dB pad at the mixer LO port to insure reasonable terminations in my IF circuit. The 160-meter transceiver I built is entirely passive (negative gain) from the antenna to the first receiver mixer, so some loss of ultimate mixer performance was permissible. The first active stage is the dual-gate MOSFET amplifier used to provide gain, matching, and filter switching. The Collins filters used (FA series) follow and were operated parallel-resonant for high impedance to match to the MOSFET *amplifier output*. The filter's output coils were coupled in the series-resonant mode to achieve a 50-ohm impedance, allowing efficient use of diode switching between the SSB and CW filters and to match the low impedance input of the MC1349 IF amplifier.[3] A simple circuit (**fig. 1**) allows simple selection of alternate bandwidth filters.

The heart of the IF stage is the MC1349 IC, an improved version of the Motorola MC1350. Besides the MOSFET preamp for the filter, it is the only other active component required for the IF stage. Notice that pin 5 is the AGC input for the device. In my unit I used audio-derived AGC with good results, but simple RF-derived AGC circuits have been published for this

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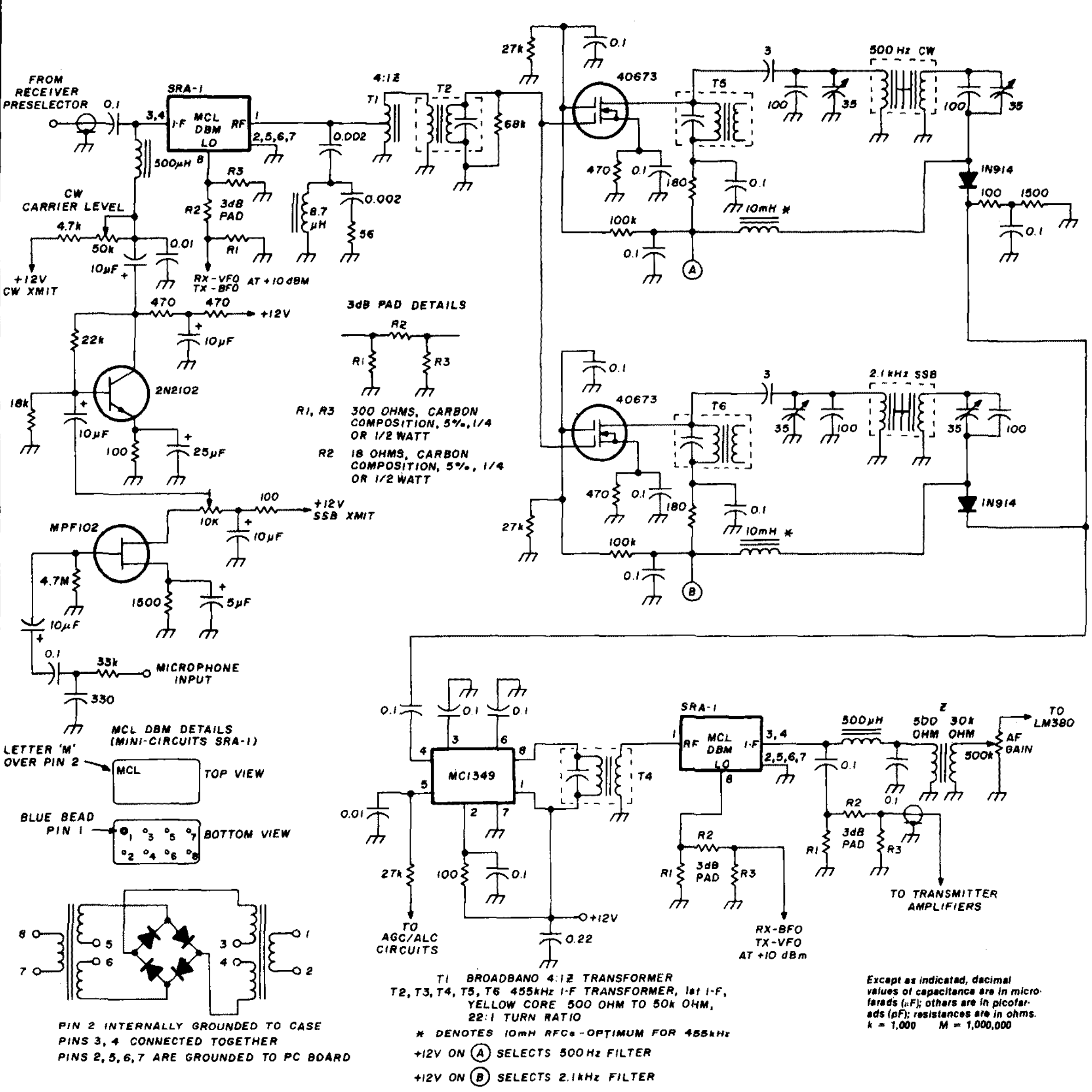

fig. 1. Quasi-bilateral IF for transceivers. Schematic shows simple method of switching between filters of different bandwidths. More filters can be added by duplicating the dual-gate FET stage for each additional filter desired. Author's transceiver incorporates one additional filter for AM reception.

device.[4] The transmitter ALC signal, also common to the AGC buss, was developed in my transceiver by diode sampling to the transmitter PA output. If an MC1350 is used, pin 3 should be grounded; for the MC1349 pin 3 should be bypassed to ground. In the MC1349, pin 3 is connected to the emitter of the input transistor and sets the device input impedance and gain. I used pin 3 in my unit to couple the IF signal to a separate AM IF stage and detector. The MC1349 output is transformer coupled to the second MCL DBM RF port. In receive this DBM serves as the product detector, with the BFO energy coupled to the LO port. Note that the recovered audio is through a three-element low-pass filter to remove any IF RF from reaching the audio stages.

The first bypass capacitor of the filter does not go directly to ground but is connected to a 3-dB attenuator. During transmit, the BFO is switched to the first DBM, and the VFO to the LO port of the second DBM. This arrangement allows the desired 160-meter prod-

uct from the second DBM to be passed onto the transmitter preselector and amplifiers with minimum losses and good termination to the mixer port.

BFO and VFO signal switching between the LO ports may be achieved through miniature relays or by using solid-state circuitry. No relays were used in my transceiver; the preselector tracked the main VFO and was switched for common duty for both the transmitter and receiver. Recovered audio from the diode ring detector was sufficient to drive a single LM380 audio IC. The AGC range is 70 dB.

Although I haven't tried it yet, the performance of this circuit at 9 MHz should be better. The newer crystal filters will have less insertion loss, and the DBMs I used were not rated for best operation below 500 kHz at the LO and IF ports. Of course the values for the diplexer[5] and for the audio coupling RF chokes would have to be changed accordingly as well as the transformers used if another IF is chosen. While the circuitry presented here could certainly stand improvement, it was my intent only to outline my solution for a novel IF arrangement. As this circuit best lends itself to a single conversion transceiver, the use of a 455-kHz IF is probably precluded above 7 MHz to avoid image problems. The FA series Collins filters I used are the same as those used in the Collins 75S series receivers; they are expensive if purchased new, but can often be found at flea markets for reasonable prices. Several dealers regularly offer imported 455-kHz mechanical filters and surplus crystal filters in the 5-9 MHz range at very attractive prices.

While no actual gain or noise figure measurements were made, the losses between the antenna and first mixer are excessive in my transceiver. The input and output of the lightly coupled preselector are diode switched for a total path loss exceeding 8 dB. For operation on 80 or 160 meters this is acceptable if a full-size antenna is used. My objective in doing this was to allow the noise floor of the receiver to be no less than required for everyday use on 160 meters, thereby improving the capability of the receiver to handle strong signals. With a carefully designed bandpass filter, such as the low loss units designed by Wes Hayward, W7ZOI, operation of this IF on any HF band should be possible without additional front end gain.[6]

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ham radio

microstrip impedance program

Use your TRS-80, HP-41, or HP-67/97 to determine stripline width and other parameters

Even though programmable calculators and computers can easily be used to perform microstrip impedance line calculations, I haven't seen many suitable programs in print.

The two synthesis programs listed here were written for use on the Radio Shack TRS-80™ computer (**fig. 1**) and the HP-67/97 or HP-41 (**fig. 2**). Both may be adapted for use on other programmable calculators and personal computers.

The programs allow the user to input a desired microstrip impedance value, frequency, relative dielectric constant and thickness to calculate the correct linewidth for that impedance. It will also output the effective dielectric constant, relative velocity factor, and 90-degree microstrip line length.

The mathematics used represent a good closed form approximation for line impedances between 20-110 ohms and will deliver results that fall within 3 percent of Wheeler's line impedance values.

Table 1 contains a list, printed out using a TRS-80, of microstrip line impedances from 10 to 100 ohms for a 0.025 *inch (6.35 mm)* thick dielectric of 10 at 2.0 GHz. It also shows the effective dielectric constant for lines, relative phase velocity, and 90 degree length.

The program listing in **fig. 1** is a short closed form approximation. No adjustments (such as a conductor thickness correction or variation of equations used at different impedance values) are made to improve accuracy. The ±2 percent accuracy is sufficient for most applications.

table 1. Microstrip line impedances from 10 to 100 ohms for a 0.025 inch (6.35 mm) thick dielectric of 10 at 2.0 GHz.

ER = 10 Thickness = 0.025 inches Frequency = 2 GHz
Enter Z_0 impedance wanted (LOW, HIGH, STEP)?
10,100,10

impedance	width (inches)	ER-effective	velocity factor	90-degree (inches)
10.0	0.2446	8.7	0.340	0.500
20.0	0.1026	7.9	0.355	0.523
30.0	0.0572	7.4	0.367	0.540
40.0	0.0355	7.1	0.376	0.553
50.0	0.0233	6.8	0.383	0.564
60.0	0.0156	6.6	0.390	0.574
70.0	0.0106	6.4	0.395	0.582
80.0	0.0072	6.3	0.400	0.589
90.0	0.0049	6.1	0.404	0.595
100.0	0.0033	6.0	0.408	0.601

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5. J. F. White, *Microwave Semiconductor Engineering*, Van Nostrand & Reinhold Publishers, 1982, (ISBN 0-442-29144-2).

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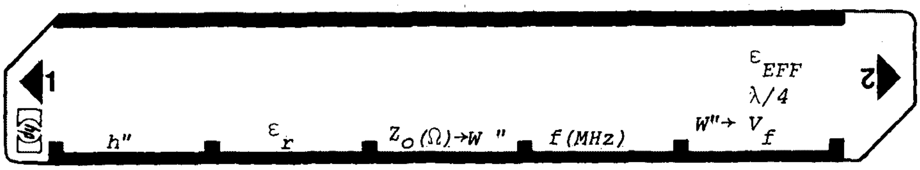
```
10 CLEAR 100
20 A$="###.#":B$=".####"
30 H2=.06:ER=4.8:ZO=50
40 CLS:PRINTCHR$(23):PRINT:PRINT:PRINT:PRINT"     MICROSTRIP CALCULATOR"
50 PRINT:PRINT
60 PRINT"      D.C. MITCHELL-K8UR"
70 FORI=0TO500:NEXT
80 CLS:PRINT"                DIELECTRIC CONSTANTS"
90 PRINT STRING$(55,"-")
100 PRINT"ER OF FIBERGLASS-EPOXY P.C.=4.8"
110 PRINT"ER OF TEFLON OR SIMILAR FLUROCARBON POLYMERS=2.5"
120 PRINT "ER OF 99% ALUMINA DIOXIDE CERAMIC = 10"
130 PRINT"THE DIELECTRIC THICKNESS (H) OF 1/16'' P.C. STOCK DOUBLE CLAD W/1 OZ./SQ.FT. COPPER=.0597"
140 PRINT"FOR 1/32'' THICK P.C.B. USE .0285";:PRINT" (DEFAULT ZO=50,ER=4.8,H=.0597)"
150 PRINT
160 INPUT"ENTER THICKNESS OF P.C.B";H2
170 INPUT"ENTER ER OF P.C.B.";ER
175 INPUT "ENTER FREQUENCY IF 90 DEG LENGTH IS DESIRED (GHZ)";FQ
180 CLS:PRINT "ER=";ER,"THICKNESS=";H2;"INCHES","FREQ=";FQ;"GHZ"
190 INPUT"ENTER ZO IMPEDANCE WANTED (LOW, HIGH, STEP)";ZL,ZH,ZS
200 PRINT "IMPEDENCE   WIDTH(IN.)     ER-EFFECTIVE       VR      90DEG.(IN)"
210 FOR ZO=ZL TO ZH STEP ZS
220 W1=H2
230 W=377*H2/(SQR(ER)*ZO*(1+1.735*ER[-.0724*(W1/H2)[-.836))
240 IF INT(W*10000)/10000=W1 THEN 270
250 W1=INT(10000*W)/10000
260 GOTO 230
270 PRINT USING A$;ZO;:PRINT "         ";:PRINTUSING B$;W1;
280 EF= ((ER+1)/2 + (ER-1)/2 * (1+10*(H2/W1))[-.5)
290 VR=1/SQR(EF)
295 L=11.785/4/FQ*VR
300 PRINT "          ";:PRINTUSING A$;EF;:PRINT "          ";: PRINT USING "#.###";VR;:PRINT"     "
310 NEXT ZO
```

fig. 1. TRS-80™ microstrip line program listing.

User Instructions



STEP	INSTRUCTIONS	INPUT DATA/UNITS	KEYS		OUTPUT DATA/UNITS
1	*Input thickness of dielectric in inches*	h"	A		
2	*Input relative dielectric constant*	ε_r	B		
3	*Input desired impedance* Z_o	Z_o Ω	C		*Width "*
4	*Input frequency in MHz*	f MHz	D		
5	*Input W, width of line in inches*	W in	E		ε_{EFF}
					$\lambda/4$ (inch)
					V_f

fig. 2A. User instructions for microstrip impedance line program for HP-67/97 or HP-41 calculators.

STEP	KEY ENTRY	KEY CODE	COMMENTS
001	f LBL A	31 25 11	
	STO 2	33 02	
	1	01	
	0	00	
	0	00	
	0	00	
	STO 3	33 03	
	•	83	
	8	08	
010	3	03	
	6	06	
	CHS	42	
	STO 5	33 05	
	•	83	
	0	00	
	7	07	
	2	02	
	4	04	
	CHS	42	
020	STO 6	33 06	
	1	01	
	•	83	
	7	07	
	3	03	
	5	05	
	STO 7	33 07	
	h RTN	35 22	
	f LBL B	31 25 12	
	STO 4	33 04	
030	h RTN	35 22	
	f LBL C	31 25 13	
	STO 0	33 00	
	RCL 2	34 02	
	STO 1	33 01	
	f LBL 0	31 25 00	
	RCL 0	34 00	
	RCL 6	34 06	
	h y^x	35 63	
	RCL 7	34 07	
040	x	71	
	RCL 1	34 01	
	RCL 2	34 02	
	÷	81	
	RCL 5	34 05	
	h y^x	35 63	
	x	71	
	1	01	
	+	61	
	RCL 0	34 00	
050	x	71	
	RCL 4	34 04	
	$\sqrt{x}$	31 54	
	x	71	
	1/x	35 62	
	RCL 2	34 02	
	x	71	

STEP	KEY ENTRY	KEY CODE	COMMENTS
	3	03	
	7	07	
	7	07	
060	x	71	
	RCL 3	34 03	
	x	71	
	f INT	31 83	
	RCL 3	34 03	
	÷	81	
	RCL 1	34 01	
	g x = y	32 51	
	R/S	84	
	h x < > y	35 52	
070	STO 1	33 01	
	GTO 0	22 00	
	R/S	84	
	f LBL D	31 25 14	
	STO 8	33 08	
	h RTN	35 22	
	f LBL E	31 25 15	
	RCL 2	34 02	
	÷	81	
	•	83	
080	6	06	
	h x < > y	35 52	
	g x > y	32 81	
	GTO 1	22 01	
	•	83	
	0	00	
	2	02	
	9	09	
	7	07	
	f COS	35 63	
090	x	71	
	GTO 2	22 02	
	f LBL 1	31 25 01	
	•	83	
	1	01	
	2	02	
	5	05	
	5	05	
	h y^x	35 63	
	•	83	
100	6	06	
	3	03	
	x	71	
	f LBL 2	31 25 02	
	RCL 4	34 04	
	1	01	
	-	51	
	x	71	
	1	01	
	+	61	
110	STO 9	33 09	
	f $\sqrt{x}$	31 54	
	h 1/x	35 62	

STEP	KEY ENTRY	KEY CODE
	1	01
	1	01
	8	08
	1	01
	1	01
	RCL 8	34 08
	÷	81
120	RCL 9	34 09
	f $\sqrt{x}$	31 54
	÷	81
	4	04
	÷	81
	RCL 9	34 09
	f - x -	31 84
	h RV	35 53
	f - x -	31 84
	h RV	35 53
130	R/S	84
140		
150		
160		

REGISTERS

0	1	2	3	4	5	6	7	8	9	A	B	C
Z_o Ω		h	1000	ε_r	-.836	-.0724	1.735					
S0	S1	S2	S3	S4	S5	S6	S7	S8	S9	D	E	I

fig. 2B. Program listing for microstrip impedance line program for HP-67/97 or HP-41 calculators.

back to basics: transistor biasing

Add a bias network to minimize the effects of temperature change

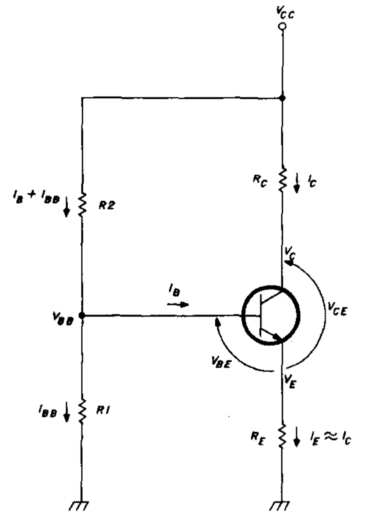
This article has been adapted, with permission, from Chris Bowick's *RF Circuit Design*, published by Howard W. Sams & Company, Indianapolis, Indiana, and available from Ham Radio's Bookstore, Greenville, New Hampshire 03048, for $25.45, postpaid.

In most RF amplifier designs, very little thought is given to the design of bias networks for the individual transistors involved. Sometimes this lack of interest is justified: if, for instance, the amplifier was expected to operate only at room temperature, there would be little need to spend much time developing an extremely temperature-stable DC operating point. If, on the other hand, the amplifier was expected to operate reliably and maintain certain specifications (gain, noise figure, etc.) over large temperature extremes — as in outdoor mounted preamplifiers, for example — the DC bias network would have to be carefully considered. A quick look at the Y and S-parameter curves for most transistors reveals that a change in the transistor's bias point does in fact change all of its RF operating characteristics. This means that the DC operating point must remain stable under specified operating conditions or the RF characteristics may change drastically.

factors affecting operating point

Two basic internal transistor characteristics — ΔV_{BE} and $\Delta\beta$ have a profound effect upon the transistor's DC operating point over temperature. The object of a good temperature-stable bias design (see fig. 1) is to minimize the effects of these parameters.

As the operating temperature *increases* the base-to-emitter voltage, V_{BE}, of a transistor *decreases* at the rate of about 2.5 mV/°C from its nominal room temperature value of 0.7 volts for a silicon device. As



step

1. Choose the operating point for the transistor:
   $I_C = 10$ mA $\quad V_C = 10$ V $\quad V_{CC} = 20$ V $\quad \beta = 50$
2. Assume a value for V_E considering bias stability (see text):
   $V_E = 2.5$ volts
3. Assume $I_E \approx I_C$ for high beta transistors
4. Calculate R_E, knowing I_E and V_E:
   $$R_E = \frac{V_E}{I_E} = \frac{2.5}{10 \times 10^{-3}} = 250 \text{ ohms}$$
5. Calculate R_C, knowing V_{CC}, V_C, and I_C:
   $$R_C = \frac{V_{CC} - V_C}{I_C} = \frac{20 - 10}{10 \times 10^{-3}} = 1000 \text{ ohms}$$
6. Calculate I_B, knowing I_C and β:
   $$I_B = \frac{I_C}{\beta} = 0.2 \text{ mA}$$
7. Calculate V_{BB}, knowing V_E and V_{BE}:
   $V_{BB} = V_E + V_{BE} = 2.5 + 0.7 = 3.2$ volts
8. Assume a value for I_{BB}, the larger the better (see text)
   $I_{BB} = 1.5$ mA
9. Calculate R1, knowing I_{BB} and V_{BB}:
   $$R1 = \frac{V_{BB}}{I_{BB}} = \frac{3.2}{1.5 \times 10^{-3}} \text{ ohms} = 2133 \text{ ohms}$$
10. Calculate R2, knowing V_{CC}, V_{BB}, I_{BB}, and I_B:
   $$R2 = \frac{V_{CC} - V_{BB}}{I_{BB} + I_B} = \frac{20 - 3.2}{1.7 \times 10^{-3}} = 9882 \text{ ohms}$$

fig. 1. Base resistor divider plus emitter resistor determines transistor's operating point.

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V_{BE} decreases, more base current is *allowed* to flow, in turn producing more collector current, which is exactly what we would like to prevent. The total change in V_{BE} for a given temperature change is called ΔV_{BE}. The primary external circuit factor that tends to minimize the effects of ΔV_{BE} and over which the circuit designer has control is the emitter voltage V_E of the transistor. A decrease in V_{BE} with temperature causes an increase in emitter current and a subsequent increase in V_E. The increase in V_E is a form of negative feedback which tends to reverse bias the base-emitter junction and, therefore, *decrease* the collector current. A decrease in V_{BE}, therefore, tends to be counteracted by the increase in V_E and the collector current does not increase as much with temperature. If these observations are put into equation form we have:

$$\Delta I_C \approx -\frac{\Delta V_{BE} I_C}{V_E} \qquad (1)$$

where ΔI_C = change in collector current
I_C = quiescent collector current
ΔV_{BE} = change in base-to-emitter voltage
V_E = quiescent emitter voltage

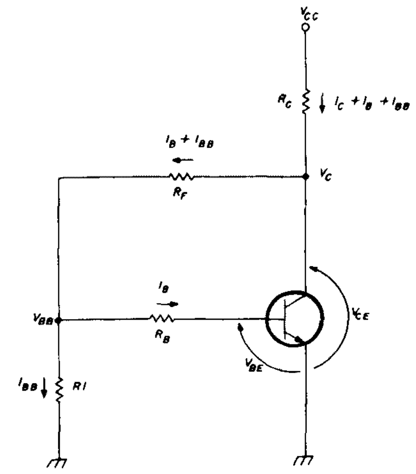
Thus, if V_E were made equal to 20 times ΔV_{BE}, then the collector current would change only 5 percent over temperature due to ΔV_{BE}. It is the value of the emitter voltage (V_E) and not the value of the emitter resistor (R_E) which is the important bias design criteria.

From eq. 1 it appears that the higher V_E is, the better. This would be true if we only had to bias the transistor at a specific operating point. Obviously, there are other things which must be considered in the design. A high emitter voltage, for instance, does tend to waste power and decrease the AC signal gain. A bypass capacitor across R_E at the signal frequency is usually used to prevent the loss in gain, but the wasted DC power may still be a problem.

If the amplifier is to operate over a temperature change not exceeding ±50°C, then an emitter voltage of 2.5 volts will provide a ±5 percent variation in I_C due to ΔV_{BE}. In fact, you will find that the majority of the transistor bias networks similar to fig. 1 will provide a value of V_E from 2 to 4 volts depending upon the values of V_{CC} and V_C chosen. Higher values are, of course, possible depending upon the degree of stability you need.

varying current gain affects operating point

The change in a transistor's DC current gain, β, over temperature is also important to the circuit designer. Any variation in β produces a corresponding change in quiescent collector current and changes the transistor's designed operating point. The β of a silicon



step

1. Choose the operating point for the transistor:
I_C = 10 mA V_C = 10 V V_{CC} = 20 V β = 50
2. Assume values of V_{BB} and I_{BB} to supply constant current I_B:
V_{BB} = 2 volts I_{BB} = 1 mA
3. Calculate I_B, knowing I_C and β:
$I_B = \frac{I_C}{\beta} = 0.2$ mA
4. Calculate R_B, knowing V_{BB}, V_{BE} = 0.7 V, and I_B:
$R_B = \frac{V_{BB} - V_{BE}}{I_B} = \frac{2 - 0.7}{0.2 \times 10^{-3}} = 6500$ ohms
5. Calculate R1, knowing V_{BB} and I_{BB}:
$R1 = \frac{V_{BB}}{I_{BB}} = \frac{2}{1 \times 10^{-3}} = 2000$ ohms
6. Calculate R_F, knowing V_{BB}, I_{BB}, I_B, and V_C:
$R_F = \frac{V_C - V_{BB}}{I_{BB} + I_B} = \frac{10 - 2}{1.2 \times 10^{-3}} = 6667$ ohms
7. Calculate R_C, knowing V_{CC}, V_C, I_C, I_B, and I_{BB}:
$R_C = \frac{V_{CC} - V_C}{I_C + I_B + I_{BB}} = \frac{20 - 10}{11.2 \times 10^{-3}} = 893$ ohms

fig. 2. Collector feedback resistor minimizes temperature effects.

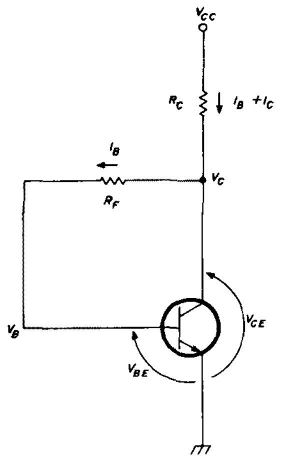
transistor typically *increases* with temperature at the rate of about 0.5 percent/°C. A ±50°C temperature variation causes β and consequently collector current to vary as much as ±25 percent.

Not only does β vary with temperature, but the manufacturing tolerance for β among transistors of the *same* part number is typically very poor. It is not uncommon, for instance, for a manufacturer to specify a 10 to 1 range for β on the data sheet (such as 50 to 500). This, of course, makes it extremely difficult to design to bias network and achieve consistent performance.

Collector current for a corresponding change in β is approximated by:

$$\Delta I_C = I_{C1}\left(\frac{\Delta\beta}{\beta_1\beta_2}\right)\left(1 + \frac{R_B}{R_E}\right) \qquad (2)$$

where I_{C1} = collector current at $\beta = \beta_1$
β_1 = lowest value of β
β_2 = highest value of β
Δ_β = $\beta_2 - \beta_1$
R_B = parallel combination of $R1$ and $R2$ in fig. 1
R_E = emitter resistor



step
1. Choose the operating point for the transistor (V_C, I_C):
   $I_C = 10$ mA $V_C = 10$ V $V_{CC} = 20$ V $\beta = 50$
2. Calculate I_B, knowing I_C and β:
   $$I_B = \frac{I_C}{\beta} = 0.2 \text{ mA}$$
3. Calculate R_F, knowing V_C, V_B, $= V_{BE} = 0.7$ V and I_B:
   $$R_F = \frac{V_C - V_B}{I_B} = \frac{10 - 0.7}{200 \times 10^{-6}} = 46.5 \text{ K}$$
4. Calculate R_C, knowing I_B, I_C, V_{CC}, and V_C:
   $$R_C = \frac{V_{CC} - V_C}{I_B + I_C} = \frac{20 - 10}{10.2 \times 10^{-3}} = 980 \text{ ohms}$$

fig. 3. Simplest bias network employs collector feedback resistor.

This equation indicates that once a transistor is specified, the only control that the designer has over the effect of β changes on collector current is through the resistor ratio R_B/R_E. *The smaller this ratio, the less the collector current varies.* Again, however, some compromise is necessary. As you decrease the ratio R_B/R_E, you also produce the undesirable effect of decreasing the current gain of the amplifier. Also, as the ratio approaches unity, the improvement in operating point stability rapidly decreases. As a practical rule of thumb for stable designs, the ratio R_B/R_E should be less than 10.

Figs. 1, 2, and **3** indicate three possible bias configurations for bipolar transistors in order of *decreasing* bias stability. Complete step-by-step design instructions and an example are also included with each circuit configuration. Note that the bias networks of **fig. 2** and **3** do not contain the emitter resistor (R_E), which provides the negative feedback to counteract collector current variations over temperature. Instead, resistor R_F is connected from the collector to the base of the transistor to provide the negative feedback. Obviously, for these two designs, the designer neither has control over the ratio R_B/R_E, nor the voltage V_E of **fig. 1**. Surprisingly, however, R_F works quite well in minimizing the effects of transistor parameter variations over temperature.

ham radio

VHF/UHF WORLD

Joe Reisert W1JR

low-noise GaAs FET technology

Over the last 20 years state-of-the-art low-noise receiver design has advanced at an incredible rate. Noise figures are now approaching 0 dB on all Amateur bands from 2 meters through 23 cm (1296 MHz), and aren't that much higher on 13 cm (2300 MHz). Most of this progress is due to advances in GaAs FET (Gallium Arsenide Field Effect Transistor) technology.

This would appear to be an appropriate time for a short historical review of the rapid changes in the state-of-the art of low-noise amplifiers. In this column, I'll also provide an overview of state-of-the-art design and present examples of some typical circuits being used to obtain extremely low noise figures. I'll also try to look a bit into the future. Because of space limitations, I'll concentrate on 2 meters through 70 cm (432 MHz), presenting straightforward designs. If you tell me you're interested, I'll cover high-performance circuit techniques applicable to use on the higher frequencies in a future column.

history

Only 25 years ago vacuum tubes such as Western Electric 417A and the gold plated 416B triodes with associated high voltage supplies and blower cooling were the state-of-the-art on the VHF bands.[1,2] Nuvistors like the 6CW4, although less expensive and easier to use, came later, but they had higher noise figures.

Early in the 1960s, paramps (parametric amplifiers) made the first two-way EME contacts possible on 23 cm and later on 70 cm.[3,4] However, these amplifiers were quite complex and generally required ferrite circulators, special diodes in a complex matching structure, and, of course, a very high frequency (typically 5 to 10 times the operational frequency of the preamp) pump oscillator. At first a klystron pump was used, but it required a regulated high voltage and was very temperature sensitive. Later (about 1970), Gunn diode oscillators were used, but they too were not for the neophyte. When the ultimate in low-noise figure was required, paramps were mounted right at the antenna, but drift and temperature changes, not to mention inconveniences in tuning, made this a last resort!

table 1. Typical low-noise GaAs FET DC and RF parameters:

DC parameters	
BV_{ds}	5 volts
BV_{gs}	−5 volts
V_p	−1 to −6 volts
I_{dss}	20-125 mA
I_d	10-50 mA
power dissipation	250-500 milliwatts
g_m	20-100 millimhos
RF parameters	
F_{max}	50-100 GHz
MAG*	15-20 dB
minimum noise figure*	0.5-2.0 dB

*At specified test frequency.

Low-noise JFETs as well as low-noise bipolar transistors became available in the late 1960s.[5,6] The really reliable low-noise UHF transistor amplifiers, however, weren't readily available until the 1970s when the FMT4575 and NE64535 arrived.[7,8] Many of these devices required special filtering or alignment techniques and were quite expensive ($30 to $55 each).

Semiconductor experts soon realized that the state-of-the-art was rapidly closing in on the maximum usable frequency where silicon bipolar transistors could operate even when using specialized techniques such as arsenic emitters. Thus when GaAs MESFETs (Metal Semiconductor FET — the full formal name for a GaAs FET) using a narrow Schottky barrier gate were first developed,[9,10] they were received with great interest.

The first commercial GaAs FET was the Fairchild Microwave FMT-900 introduced in 1972, at $500 each. It was good news to Amateurs when affordable low-noise GaAs FETs became available in the 1970s;[11] they were soon in widespread use on 70 cm and later 2-meter EME. Now they're used almost exclusively on EME from 2 meters through 13 cm. Many VHF/UHFers are presently using inexpensive (under $6) single and dual-gate GaAs FETs, even for routine weak signal and tropo propagation work.

GaAs FET parameters

What are these wonderous GaAs FETs anyhow? In many ways the single-gate GaAs FET operates like a triode vacuum tube, but with extremely low voltage (typically less than 5 volts) and no filaments to heat! There are some similarities to ordinary FETs. The input is usually the gate, which has a high input impedance (1-10 kilohms typical at VHF/UHF). The common element is usually the source and the drain is the output, which has a typical impedance in the region of 200 to 1000 ohms. The width of the gate, typically 0.5 to 2 microns, has the greatest effect on gain and maximum frequency of operation, with the narrower widths required for the higher frequencies (more on this later — see **fig. 1**).

Table 1 shows that each GaAs FET device has certain important low frequency or DC parameters such as: BV_{ds}, the maximum drain to source breakdown voltage; BV_{gs}, maximum gate to source breakdown voltage; V_p, pinchoff voltage, the minimum gate voltage required to decrease drain current to a specified low value; I_{dss}, maximum drain current with zero volts gate bias; I_d, typical operating drain current; maximum power dissipation, and g_m or transconductance (in millimhos).

Table 1 also shows the most important GaAs FET RF characteristics such as F_{max}, MAG, NF_{min}. F_{max} is the maximum frequency of oscillation. The MAG (maximum available gain) is usually specified by the manufacturer at only one or two frequencies: 10 to 25 percent of F_{max} at a specified drain current, typically at 25 to 50 percent of I_{dss}. Noise figure is usually specified at the same frequency or frequencies at a drain current of 10 to 20 percent of I_{dss}.

One of the important properties of the GaAs FET is that noise figure is primarily limited by thermal noise generated in the channel or gate of the device. Plots of maximum gain and lowest possible noise figure over the usable frequency spectrum are often provided on data sheets. As with bipolar and JFET devices, the gain increases with decreasing frequency while noise figure may decrease slightly below the test frequency. Typical input and output impedances over a wide frequency range are often shown on the GaAs FET data sheets as well as the optimum source impedance for minimum noise figure.

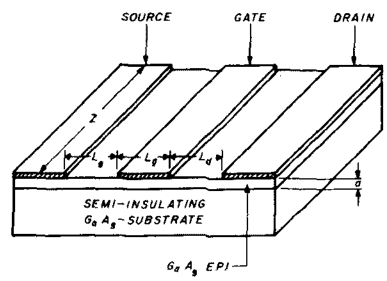


fig. 1. Cross-section of a GaAs FET.

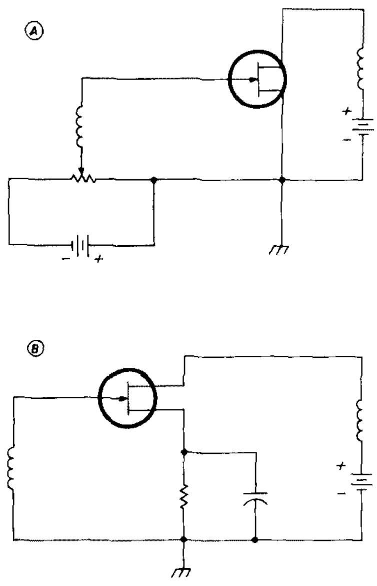


fig. 2. Typical GaAs FET biasing schemes: (*A*) battery bias; (*B*) source bias.

Recently, dual-gate GaAs FETs have been manufactured primarily for the low-cost home entertainment market. These devices are often used as amplifiers in the cascode configuration with optional AGC applied to the second gate. They've also been used as mixers, with the local oscillator applied to gate 2 in a manner similar to the dual-gate MOSFETs in the HF and VHF spectrum. SHF dual-gate GaAs FETs are also becoming available, but are still quite expensive — typically over $50.

So far I've described the so-called low-noise or small-signal GaAs FETs, primarily used in preamplifiers. High-power GaAs FETs are available with up to 8 watts of linear power; some of these also perform well in low-noise preamplifiers (more on this later). These devices also have potential for very high dynamic range. GaAs FETs are already appearing in commercial transmitter chains as well as in receivers!

A partial list of the most popular GaAs FETs typically used by Amateurs is shown in **table 2**. Typical noise figures and cost are also listed for comparison. Furthermore, new and improved devices are constantly becoming available as the technology matures. Low noise GaAs FETs usable to well beyond the 3 cm (10 GHz) band are now available, but are still quite expensive — over $50.

biasing GaAs FETs

GaAs FET biasing is quite different from bipolar transistors. In fact, the GaAs FET is more like the old triode vacuum tube. The commmon source configuration is presently the most popular topography. For proper operating parameters, the gate must be reverse biased with respect to the source.

When GaAs FETs were first introduced, battery bias was usually applied to the gate as shown in **fig. 2A**.[11] Many users, however, experienced problems such as burnouts due to battery failure or poor contacts on battery holders. Later, commercial amplifier manufacturers developed exotic bias supplies.[12] Source biasing with a single resistor (**fig. 2B**) became popular once it was determined that the source could be adequately RF bypassed without lowering gain or circuit performance.[13]

input impedance matching

GaAs FETs have much higher input impedances than bipolar transistors. As a result, different input matching techniques are required. The most popular are the π-network, the broadband inductor match, and the tank circuit (**fig. 3**).

In the 1960s, π-network input matching became popular, with low-noise bipolar transistors used on 23 cm. Though Amateur GaAs FET preamplifiers on 23 and 13 cm often used this technique, shown in **fig. 3A**,[14] it never became very popular in Amateur preamplifiers for 2 meters through 70 cm.

Several years ago the NRAO (National Radio Astronomy Observatory) developed cooled GaAs FET preamplifiers for radio astronomy.[15] These used a single inductor to match to the gate and reactive lossless feedback in the source (**fig. 3B**). This technique had previously been used by Reisert in a 70-cm bipolar amplifier;[16] it has since been used by Angle and Sutherland in Amateur GaAs FET preamplifiers.[17,18] However, *it's more popular on 33 cm* (902 MHz) and above, since high-pass filtering (which can create additional losses) is required ahead of the preamplifier.[19]

The most popular input matching technique used by Amateurs on 2 meters through 70 cm is probably the tank circuit. The tapped inductor version is the least expensive and potentially has the lowest loss, since only one tuning capacitor is required (**fig. 3C**). The capacitance coupled tank circuit (**fig. 3D**), however, is even more commonly used because it doesn't require the tedious selection of the proper tap point.

The tank circuit in **fig. 3D** can match a GaAs FET over a very wide impedance range from optimum gain to optimum noise figure. In addition, it has built-in selectivity, eliminating the need for additional external filtering. However, the unloaded *Q* (more on this later) of all the components in the network must be very high if the inherent low-noise figure of the device is to be achieved.

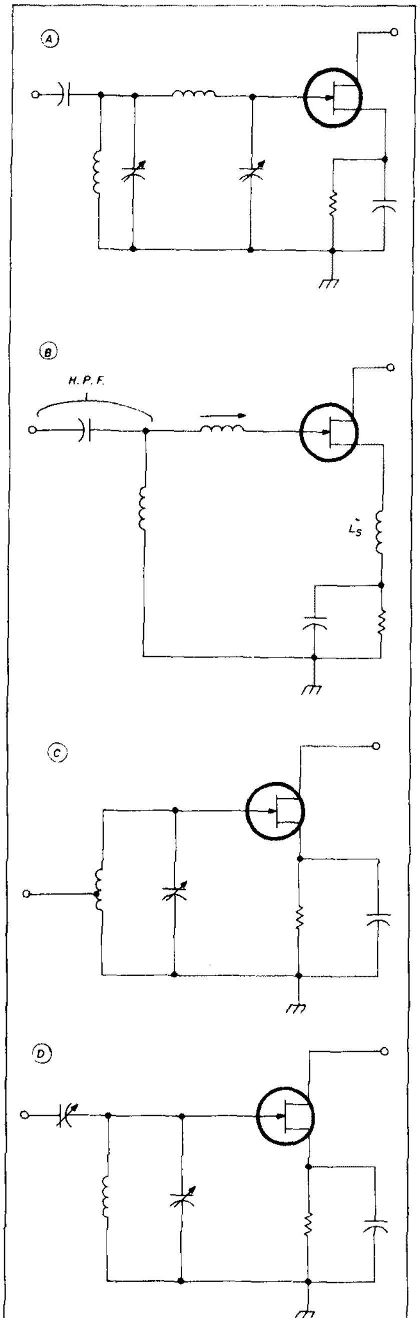


fig. 3. Typical input matching schemes for GaAs FET preamplifiers: (*A*) π-network; (*B*) broadband inductor; (*C*) inductively coupled tank circuit; (*D*) capacitance coupled tank circuit.

output impedance matching

Finally, the GaAs FET must be matched to the next stage. Here again, special techniques are required. The output impedance of most small-signal GaAs FETs is moderate (100-1000 ohms typically). Furthermore, high gain increases the chances of instability or oscillations. Therefore, the earlier Amateur GaAs FET circuits used simple loading techniques such as those shown in **fig. 4A**.[7,11] Sando used tuned tank circuits (**fig. 4B**)[20,21] and Sutherland advocated the use of a broadband 4:1 bifilar wound transformer (**fig. 4C**).[22]

All of these techniques have advantages and disadvantages. While the resistor loading approach is simple, improves stability, and is inexpensive, it lowers gain and output power by up to 3 dB. The tuned circuit delivers more gain but is more expensive and requires more input to output circuit isolation to prevent instability. The 4:1 transformer is easy to construct but has slightly less gain as well as poor to moderate output VSWR.

recommended circuit

Figure 5 shows a recommended GaAs FET circuit for operation on 2 meters through 70 cm. The input matching is the capacitance coupled tank circuit (**fig. 3D**) since it permits either conjugate matching for low input VSWR and maximum gain or optimum source impedance for minimum noise figure. The output circuit uses a 4:1 transformer as discussed earlier. One transformer covers the entire operating range from 2 meters to 70 cm. Notice that a ferrite bead is also placed on the drain lead of the GaAs FET. This reduces gain slightly 0.5-0.75 dB maximum, but improves stability, especially in the microwave region where oscillations often occur when using GaAs FETs.

Source biasing is used because it is inexpensive and self-protecting. Most single-gate GaAs FETs have two source leads and therefore lend themselves to using two source bypass capacitors and resistors. This also simplifies bias resistor selection and increases device protection in case one of the source leads should become loose. Ferrite beads are also slipped on

the source resistors since they've been reported to prevent certain instabilities in particularly stubborn circuits.

Drain voltage is supplied by a three-terminal voltage regulator that acts like a zener diode but operates over a much wider range of device current. It is followed by a resistor and zener diode for over-voltage/current protection. Placing the resistor after the regulator protects the zener and decreases the drain voltage if too much drain current is present. For the ultimate in low-noise figure, this resistor value can be raised or lowered slightly for optimum drain voltage, which is typically not critical.

dual-gate GaAs FET circuit

A typical dual-gate GaAs FET circuit is shown in fig. **6**. This is virtually the same circuit as the one just discussed but with an additional voltage source for gate 2. Most VHF/UHF dual-gate devices are very low in cost because they're primarily aimed at the TV and consumer market and have often equalled the performance of their single gate cousins, especially on 2 meters through 70 cm.[23] They typically run at higher I_d.

component selection

One of the first decisions you'll have to make is which GaAs FET to use. Some GaAs FETs (with typical specifications and prices) popular with Amateurs, are shown in **table 2**. The beauty of the circuit diagrams shown in **figs. 5** and **6** is that they'll work well through 500 MHz with just about any GaAs FET as long as the input inductor, L1, (see the parts list on **figs. 5** and **6**) is changed for the proper frequency range.

When striving for a very low-noise figure below 500 MHz, don't use a high-priced GaAs FET. The narrow gate (less than 1 micron) higher frequency units will rarely yield any better performance at lower frequencies than the less expensive units and will frequently be more susceptible to instabilities.

It has been speculated that 1/f noise, the property of a solid-state device to have increased noise figure when operated below a certain low frequency cutoff, can actually increase noise figure at the lower VHF frequencies if a device with too high an F_{max} is used. Furthermore, even the low-cost, lower frequency GaAs FETs have noise figures that equal the higher frequency higher cost units at *lower* frequencies. These points are illustrated in fig. **7**.

Input circuit losses must be kept to an absolute minimum as discussed in reference 19. The input circuit shown in **figs. 5** and **6** was chosen because it has such a wide range of tuning, good out-of-band rejection, and a reasonably low loaded "*Q*." But it also uses two capacitors, which will increase the dissipation loss in the tank circuit. High-*Q*, low-inductance air variable capacitors such as the Johanson 5200 series or equivalent are a

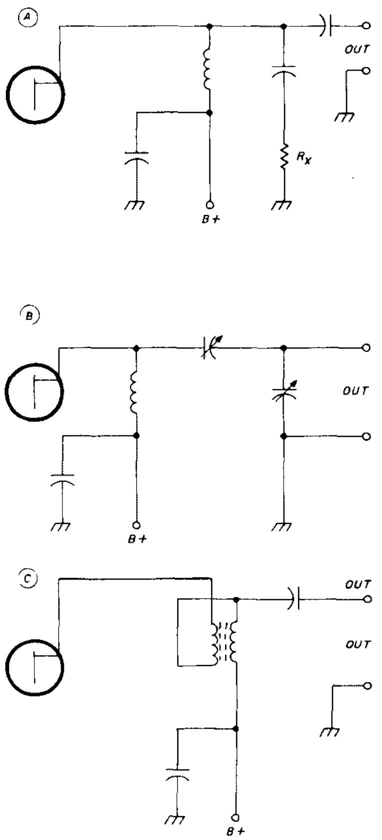
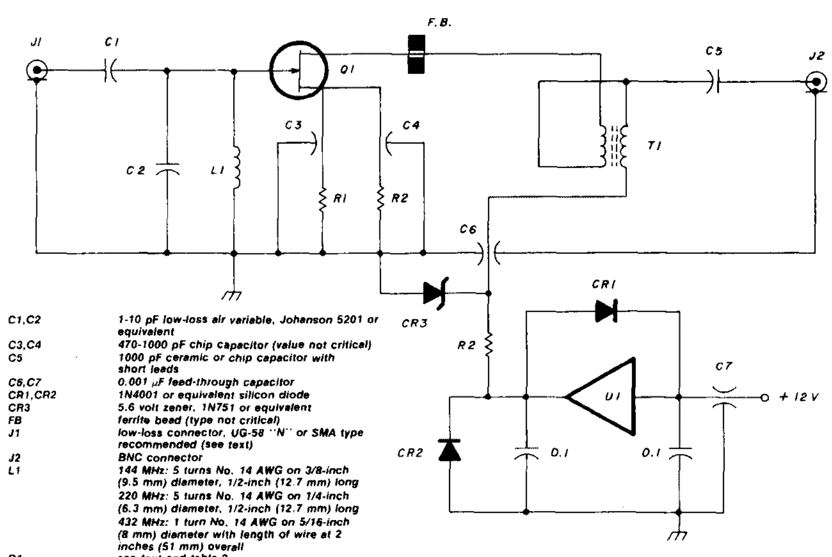
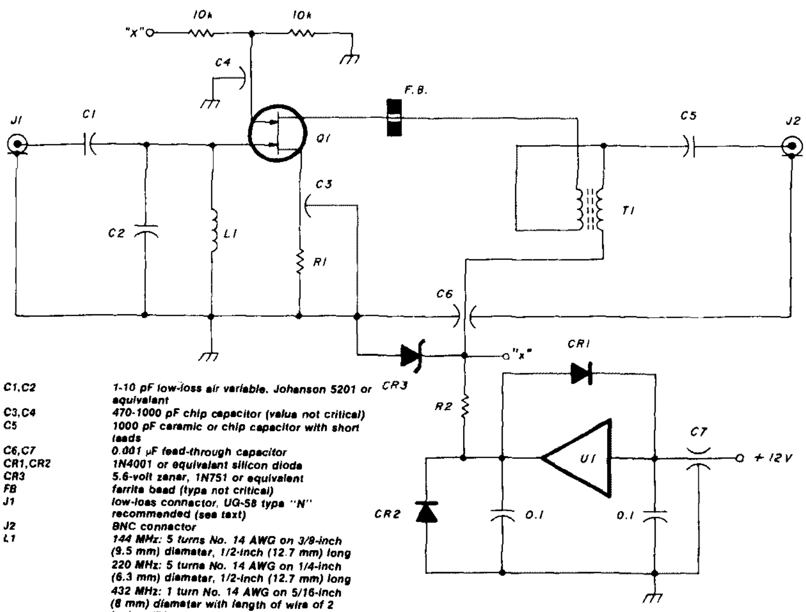


fig. 4. Typical output matching schemes for GaAs FET preamplifiers. (*A*) resistor loading; (*B*) tuned circuit; (*C*) broadband transformer.



C1,C2	1-10 pF low-loss air variable, Johanson 5201 or equivalent
C3,C4	470-1000 pF chip capacitor (value not critical)
C5	1000 pF ceramic or chip capacitor with short leads
C6,C7	0.001 μF feed-through capacitor
CR1,CR2	1N4001 or equivalent silicon diode
CR3	5.6 volt zener, 1N751 or equivalent
FB	ferrite bead (type not critical)
J1	low-loss connector, UG-58 "N" or SMA type recommended (see text)
J2	BNC connector
L1	144 MHz: 5 turns No. 14 AWG on 3/8-inch (9.5 mm) diameter, 1/2-inch (12.7 mm) long 220 MHz: 5 turns No. 14 AWG on 1/4-inch (6.3 mm) diameter, 1/2-inch (12.7 mm) long 432 MHz: 1 turn No. 14 AWG on 5/16-inch (8 mm) diameter with length of wire at 2 inches (51 mm) overall
Q1	see text and table 2
R1,R2	select for proper I_d, 300 ohms typical (see text)
R3	100 ohms typical (see text)
T1	3 turns No. 30 or No. 32 AWG bifilar wire wound on a ferrite bead
U1	5 volt, 3-terminal regulator, 78L05 or equivalent

fig. 5. Recommended GaAs FET preamplifier circuit for 144, 220, and 432 MHz operation.



C1,C2	1-10 pF low-loss air variable. Johanson 5201 or equivalent
C3,C4	470-1000 pF chip capacitor (value not critical)
C5	1000 pF ceramic or chip capacitor with short leads
C6,C7	0.001 μF feed-through capacitor
CR1,CR2	1N4001 or equivalent silicon diode
CR3	5.6-volt zener, 1N751 or equivalent
FB	ferrite bead (type not critical)
J1	low-loss connector, UG-58 type "N" recommended (see text)
J2	BNC connector
L1	144 MHz: 5 turns No. 14 AWG on 3/8-inch (9.5 mm) diameter, 1/2-inch (12.7 mm) long 220 MHz: 5 turns No. 14 AWG on 1/4-inch (6.3 mm) diameter, 1/2-inch (12.7 mm) long 432 MHz: 1 turn No. 14 AWG on 5/16-inch (8 mm) diameter with length of wire of 2 inches (51 mm) overall
Q1	see text and table 2
R1	select for proper I_d, 100 ohms typical (see text)
R2	100 ohms typical (see text)
T1	3 turns No. 30 or No. 32 AWG bifilar wound on a ferrite bead
U1	5-volt 3-terminal regulator, 78L05 or equivalent

fig. 6. Typical dual-gate GaAs FET circuit.

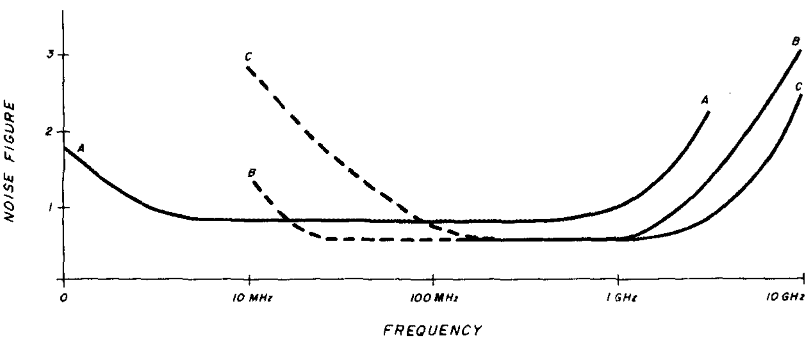


fig. 7. Typical noise figures versus frequency of several types of devices: (*A*) typical bipolar transistor; (*B*) typical GaAs FET; (*C*) narrow gate high frequency GaAs FET.

necessity if low-noise figure (operation) is required.

From laboratory measurements I've made, the best unloaded *Q* for properly designed discrete inductors is about 400-500. **Table 3** shows that the typical loaded *Q* for 2-meter through 70-cm GaAs FET preamplifiers tuned for minimum noise figure is between 8.6 and 16. Hence the calculated tank circuit losses can easily be from 0.15 to 0.35 dB with properly adjusted preamplifiers.[19] Consequently the tank circuit losses ahead of the typical GaAs FETs presently used by Amateurs are undoubtedly the main contributor to the overall noise figure and deserve lots of attention by those striving for the ultimate low-noise preamplifier for EME. The type of input connector chosen is very important, especially on 70 cm and higher. A type "N" is highly recommended. SMA connectors, although more expensive, are also an excellent choice, especially if size is a constraint.

In laboratory measurements I've made, BNC connectors often caused the noise figure to increase or become erratic, especially when you press against the side of the connector. Hence they're not recommended for use on the input to a low-noise preamplifer, since losses and VSWR translate directly to increased noise figure. However, these connectors are acceptable on the output of a preamplifier because the gain ahead of the connector will mask any possible loss.

construction techniques

Needless to say, extremely low-noise circuits such as those just described require particular attention to lead length, layout, and quality of components if the specified performance is to be obtained. Therefore I've shown a recommended construction technique in **fig. 8**. This configuration is a combination of techniques I've seen used including those of the Michigan Microwave Group and Krauss in his 13 cm preamps.[24]

A medium-size cast box such as the Bud model CU-124 or equivalent is highly recommended. Smaller boxes such as the Bud CU-123 and the Pomona Electronics 2417 have sometimes caused oscillations. Smaller boxes may require placing the input inductors too close to the DC biasing component, causing excessive coupling between them and consequently lowering the unloaded *Q*. Furthermore, unloaded inductor *Q* is almost always higher when coils are physically large and the volume surrounding them is even larger. This is the technique used to design high-*Q* helical inductors.[19]

Finally, grounding is important. I use a double-clad PC board attached to the cover of the cast box by the input and output connectors. But because typi-

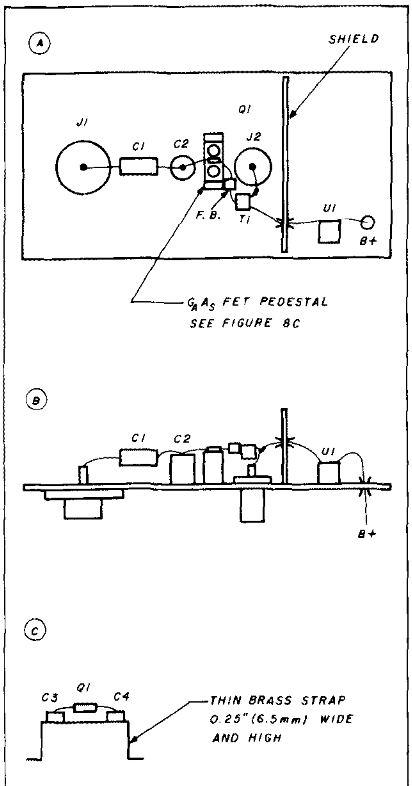


fig. 8. Recommended circuit layout for the preamplifiers shown in *figs. 5* and *6*.

cal air-variable capacitors will not go completely through the box cover and PC board, a hole large enough to pass the nut on the shunt capacitor, C2, must be drilled through the cast box. For proper grounding and lowest loss, drill two holes adjacent to this capacitor and tighten the PC board to the cover of the cast box with two short screws. A small hole can be drilled in the box cover between J1 and C2 to facilitate adjusting C1.

handling GaAs FETs

GaAs FETs with their low breakdown voltages and high impedances are far more likely to be destroyed by improper handling than their bipolar cousins. When installing a GaAs FET in a circuit, be sure your soldering iron has a grounded tip. If you're not absolutely certain the tip is grounded, place a wire with a pair of large alligator clips between the soldering iron tip and chassis ground. Touch the chassis with one hand and then pick up the GaAs FET container with your other hand. Then remove the device from its packing material while still holding on to the chassis. This will prevent static discharge when the device comes in contact with the circuit. *Caution: Don't wear woolen clothing or walk across a carpet just before working on the installation of a GaAs FET. Static electricity present can cause burnout!**

protecting GaAs FETs from RF and DC

If you use the biasing scheme recommended in **figs. 5** and **6**, there should be little chance of burnout due to DC parameters because the circuit is self limiting. Just make sure to observe the power dissipation rating when testing for I_{dss}. For ultimate DC reliability, the power supply should be of the regulated type despite the use of a three-terminal voltage regulator within the preamplifier. Diodes CR1 and CR2 in **figs. 5** and **6** provide additional protection from voltage spikes or reverse polarity. *Voltage should never be obtained from a power supply that is also used to power relays because voltage spikes may be induced by the relay coil inductance and hence cause solid-state devices to burn out. In addition, turning the power supply on and off between receive and transmit can actually increase the chances of burnout, since most solid-state devices are less likely to burn out when operating at their proper bias levels.*

The usual RF handling precautions apply. For maximum protection, the RF input level should never exceed 100 and preferably be no greater than 10 milliwatts. A dual relay protection scheme is highly recommended.[25] While GaAs FETs are not nearly as RF fragile as many persons would have you believe, they may, *when overstressed* with RF, tend to experience catastrophic failures, rendering them completely inoperative. Bipolars, on the other hand, may degrade slowly without you necessarily being aware of their progressive degradation.[26]

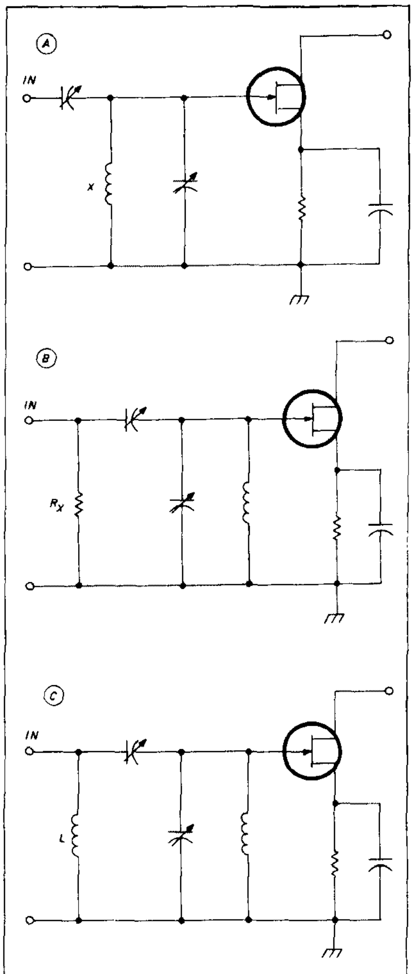


fig. 9. Some recommended protection techniques for decreasing the chances of lightning or static damage to GaAs FET preamplifiers: (*A*) capacitance coupled input tank; (*B*) resistor bleeder; (*C*) shunt inductor.

Most of the energy in static and lightning is concentrated in or below the HF spectrum. Therefore, use of the capacitance coupled tank just recommended will lessen chances of burnout since the input capacitor will act like a high-pass filter. Using a bleeder resistor or inductor across the input connector may also help, but can degrade noise figure slightly if not properly chosen (**fig. 9**). Back-to-back diodes across the input connector are not recommended because any presence

*See K4KEF's "Static Electricity and Modern Integrated Circuits," *ham radio*, March, 1984, page 33. — **Editor**

table 2. GaAs FETs presently in Amateur use.

type	NF (typically) and frequency	approximate price	notes
ALF 1023	1.3 at 4 GHz	$15.00	discontinued
D432	1.8 at 4 GHz	26.00	Dexcel
MGF 1100	2.5 at 4 GHz	7.35	dual-gate
MGF 1202	2.0 at 4 GHz	9.70	low cost MGF 1402
MGF 1402 (2SK274)	1.1 at 4 GHz	14.00	very popular
MGF 1404	0.65 at 4 GHz	66.60	expensive
MGF 1412 (2SK275)	1.0 at 4 GHz	21.00	popular above 70 cm
MRF 966	1.2 at 1 GHz	3.75	new dual-gate
NE41137 (3SK124)	1.3 at 900 MHz	5.00	dual-gate
NE72089	1.0 at 4 GHz	12.00	low-cost NE21889
3SK97	1.3 at 1 GHz	5.00	process problems
3SK121	1.5 at 800 MHz	3.50	dual-gate

of RF at lower frequencies may cause severe intermodulation products or spurious signals to appear.

testing and tuning the preamplifier

After the preamplifier is completely wired, inspect for any possible wiring errors. Then test for proper DC operation with a variable voltage power supply, first connecting the preamplifier ground to the power supply ground and then connecting the DC input and bringing the voltage up slowly to verify that the three-terminal voltage regulator is properly functioning. Now check for proper GaAs FET biasing. First connect a suitable 50 ohm termination on the preamplifier input and output connectors. Turn on the power supply and apply 12 volts to the preamplifier. In the GaAs FET, properly performing DC conditions are satisfied when the voltage drop across the 100-ohm resistor in series with the drain line is between 0.5 and 2.5 volts, indicating an I_d of 5-25 milliamperes, respectively.

The optimum value of I_d for low noise is not critical; 10-15 milliamperes is typical for low-noise GaAs FETs. But if you want to determine the optimum value, temporarily short the source lead to ground with a piece of wire and read the voltage across the 100-ohm resistor. Then calculate I_{dss}. Optimum I_d is typically 15 percent of I_{dss}. Therefore, to provide the optimum I_d for best noise figure, change the value of the source resistors until the correct I_d is obtained for each individual GaAs FET.

Next, tune the preamplifier for maximum gain, using a suitable signal source or by listening to a local station. 15-25 dB gain should be readily obtained, depending on the device and frequency range selected. Then, if you have a noise-figure generator or can obtain the use of one, optimize the input circuit for minimum noise figure. Obtaining the best noise figure will require the series input capacitance to be increased somewhat over the amount required for maximum gain. Note that the gain will drop somewhat when the circuit is optimized for best noise figure as indicated in **table 3**.

Finally, carefully measure the forward gain. If possible, also measure the reverse loss. For complete stability, the reverse loss should be at least 5 to 6 dB greater than the forward gain. For example, referring to the test data in **table 3**, the stability margin (the reverse loss minus the forward gain) at 2 meters is 9 dB for the low-noise unit but only 1.6 dB in the maximum gain unit, indicating that there are potential stability problems. A look at the 220 MHz test data will show an even worse example!

performance

Table 3 can be used as a guide for all GaAs FET preamplifier performance parameters except noise figure. As previously pointed out, the data was taken on a sample GaAs FET chosen not for lowest noise figure, but only for comparison of typical performance. Lower noise figure devices such as those shown in **table 2** should yield performance as indicated in the table.

When a GaAs FET is tuned for best noise figure, the gain will drop a few dB. The gain peak may move to a different frequency. However, this should still be more than sufficient gain even in the most stringent system. Because the suggested output circuit is a broadband device, the overall bandwidth measured for the circuit should be that of the input tank circuit. Hence, circuit losses can be easily calculated as shown in reference 19.

lower noise figures

The circuits shown should be more than adequate to yield state-of-the-art performance in the 2 meter through 70-cm spectrum. Furthermore, noise figures of less than 0.5 dB should be about all that can be used even on 70-cm EME, where the sky has a low noise temperature. Lower noise figures can sometimes be obtained by using the higher cost devices as discussed earlier. However, to lower noise figure further usually requires additional work on the components and structures used in the input tank circuit.

Q tests were performed on the microstrip circuit suggested by Sutherland.[22] However, tests typically yielded an unloaded *Q* of 400-500, no higher than the inductors suggested earlier. Although I have not tried it, adding a three-sided 1 inch (25.4 mm) square

table 3. Typical preamplifier performance: data taken with the same GaAs FET in the same circuit (fig. 5). Only the input inductor was changed for each frequency range. The GaAs FET used was not chosen for superior performance but for continuity of data. However, the data indicates the typical performance to be expected when optimizing for either gain or noise figure.

frequency (MHz)	144 (1)	144 (2)	220 (1)	220 (2)	432 (1)	432 (2)
forward gain in dB	25.4	21.00	25.20	20.10	19.20	17.70
reverse gain in dB	−27.0	−30.00	−24.00	−28.00	−25.50	−28.00
stability margin in dB	1.6	9.00	−1.20	7.90	6.30	10.30
input VSWR	1:1	4.8:1	1:1	4.2:1	1:1	3:1
output VSWR	12:1	2.6:1	16:1	1.8:1	2.2:1	1.7:1
output power at compression (dBm)	8.0	6.00	6.30	7.00	4.50	5.00
noise figure in dB	1.4	0.53	1.35	0.66	1.28	1.00
input impedance in kilohms	6.8	1.80	6.20	1.80	1.90	1.23
bandwidth in MHz	2.9	9.00	5.80	25.40	22.30	30.80
overall "Q" of amplifier	49.7	16.00	37.90	8.70	19.40	14.00

Note 1: Preamplifier gain matched.
Note 2: Preamplifier noise figure optimized.

copper shield to completely enclose the microstrip may decrease losses, since all the current surrounding the inductor would be within the shield and not circulating in the aluminum of the shielded box. Experiments were conducted using silver plated 1/4-wavelength cavities with an unloaded *Q* of over 1000. These are much larger and more expensive structures but they could lower input losses by at least 0.1 to 0.25 dB.

The air-variable capacitors used also contribute to loss. The ones suggested have a specified unloaded *Q* of 5000 at maximum capacitance at 100 MHz and commensurately lower *Q* at increasing frequency. Older models such as the Johanson 2950 series and its equivalent have an unloaded *Q* of 2500 respectively. Because the series matching capacitor seldom exceeds 3.0 pF in the 220 MHz and higher frequency preamplifiers, a physically smaller, higher *Q* capacitor such as the Johanson 5800 series with a 0.3-3 pF range and unloaded *Q* of 7500 at 100 MHz may be a way to decrease input losses further. The Johanson Giga-trim™ types are not recommended since they have lower *Q* than the air-variable types.

Finally, GaAs FETs can be cooled. Thermo-electric coolers, used by commercial manufacturers to cool GaAs FET type preamplifiers and thereby decrease noise figure, are now available to Amateurs. Liquid nitrogen has been used by some Amateurs, but is costly and requires special handling and mechanical arrangements.

looking ahead

GaAs FETs are becoming increasingly less expensive and easier to use. Almost monthly, a new lower noise figure and lower cost device seems to appear. This is particularly true as you go higher in frequency. Before long, receivers using only diode mixers will be a thing of the past, expecially at 3 cm (10 GHz), where devices with noise figures under 2 dB are now affordable. The increase in performance should be phenomenal, since most work on this band has been conducted with noise figures of 5 to 10 dB. The possibilities for SHF EME are considerable — if the industry also shifts to the production of inexpensive kilowatt transmitting devices!

The use of monolithic hybrid matched amplifiers will take hold. At the present state-of-the-art, the prices are acceptable ($5 to $20), but the silicon devices available often have noise figures of 3 to 6 dB and the lower noise devices seldom have adequate dynamic range. The present frequency limit is typically 3 GHz. However, monolithic GaAs FET hybrids are now commercially available. They hold great promise because while their price will be competitive with present silicon technology, they will go higher in frequency and have much lower noise figures and potentially higher dynamic range.

Although this is primarily a receiver preamplifier article, I should mention that power GaAs FETs are being developed for transmitter applications. They're presently available with up to 8 watts and are closing in on the 3 cm area. Soon the last bastions of tubes will be at stake.

This article was not intended to plow new ground, but instead to bring readers up-to-date on the state-of-the-art in low noise figure preamplifier design using GaAs FETs below 500 MHz. An attempt was made to take the fear out of using GaAs FETs even in your everyday circuits; if you follow the cautions mentioned above, you shouldn't have any problems.

While usually somewhat behind the commercial manufacturers, Amateurs are constantly striving to stay abreast of technology. Right now we're running neck-and-neck with industry on VHF/UHF noise figures. We have one great advantage in that we frequently only need narrow-band circuits that don't require economical reproduction. We can therefore take advantage of all the "tricks of the trade" and optimize to our hearts' content. The typical noise figures measured at VHF/UHF conferences on typical Amateur-built low-noise preamplifiers attest to the fact that we are very close to noise figure *perfection* on the VHF and lower UHF frequencies!

acknowledgements

I'd like to particularly thank Bob Sutherland, W6PO, and Shigeru Sando, JH1BRY, who encouraged me to try a GaAs FET preamp on my 70-cm EME station. I've been a firm believer in GaAs FETs ever since!

postscript

It seems hard to believe that a year has already passed since my first column in this series. I've tried to cover a lot of ground and still have plans for a few more articles dealing with important background material. When this material is all in place, I expect to buiid on this material and move forward. Your ideas for future columns are always welcome!

I sincerely want to thank in particular the *ham radio* staff — K2RR, KA1LBO, and N1ACH, who have all helped me through this year. I'd also like to thank all of you who have encouraged me with your notes and calls. Believe me, it takes many hours and a big slice out of your operating time to write this quantity of material month after month. A kind word here or there does push me along.

Unfortunately, I haven't always been able to answer all the letters received, expecially those without SASEs or those that hit me with multiple questions and new design requirements. If I were to answer all the letters as received, there'd be little time left to write the column!

Don't let this discourage you from writing. If you need an answer to a question, or if a problem occurs with something mentioned in one of my columns, please send a letter with an SASE. Keep your question brief so that I can respond quickly and easily. I'd rather answer several letters from one individual with one question per letter than multiple questions in a single letter. I hope you understand. Thank you, in advance, for your understanding and encouragement! It is truly appreciated.

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coming VHF/UHF events:

December 13, 0048 UTC: *predicted peak of Geminids meteor shower*

December 18: *EME perigee*

December 21, 1600 UTC: *predicted peak of Ursids meteor shower*

December 22, (±2 weeks): *annual secondary peak of sporadic "E" propagation*

ham radio

the ham notebook

a different approach to the inductance equation

Although most Amateurs are familiar with the well-known Wheeler equation for inductance,[1] and with its expression in another form,[2] neither is readily solvable when you want to find the number of turns required when you already know the inductance desired. Actually, in almost all Amateur applications, you first determine the inductance you need and then, with whatever wire and coil form you have available, you solve for the required number of turns. The following equation, derived from the Wheeler equation, allows you to do just that:

$$n = \frac{\frac{20L}{t} + \sqrt{\frac{400\ L^2}{t^2} + 18\ d^3L}}{d^2}$$

where n = number of turns
L = desired inductance in microhenries
d = coil diameter in inches
t = number of turns per inch (close or space wound)
t = 1/wire diameter, for close wound coils

example

d = 2.5 inches
t = 19 turns per inch
L = 145 microhenries

The number of turns will come out to be 65.

Although at first glance the equation looks formidable, it really isn't. Using a small calculator, it can be solved in a few minutes. And as the equation shown is derived from the Wheeler equation, it naturally has the same constraints as to accuracy.*

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William Vissers, K4KI

*The familiar Wheeler equation that Bill referred to, that appears in reference 2 is reproduced below:

$$L_{uh} = \frac{a^2n^2}{9a + 10b}$$

where L = inductance in microhenries
a = coil radius in inches
b = coil length in inches

The practical difficulty in applying the more commonly used equation is that wire size (diameter) has to be carefully considered. If it is not, contradictory values of *b* will be obtained. Bill's developed equation eliminates that problem. And as stated, his equation is correct for both closely and widely spaced wound coils. — **Ed.**

ham radio

simple filter tester

I build all of my projects using a minimum of simple homebrew test gear. Generally, all that's necessary is a grid-dip oscillator, signal generator, VTVM, and an inductance bridge.

Because my VTVM with an RF probe isn't sensitive enough to pick up weak signals for aligning IFs or checking crystal filter frequencies, I built a filter tester to provide this capability (see **fig. 1**).

To use, simply feed in a modulated signal (tuned to the desired frequency) from a signal generator (start with 2 volts output) to the IF strip or filter, then listen for the maximum signal output on the tester.

Ed Marriner, W6XM

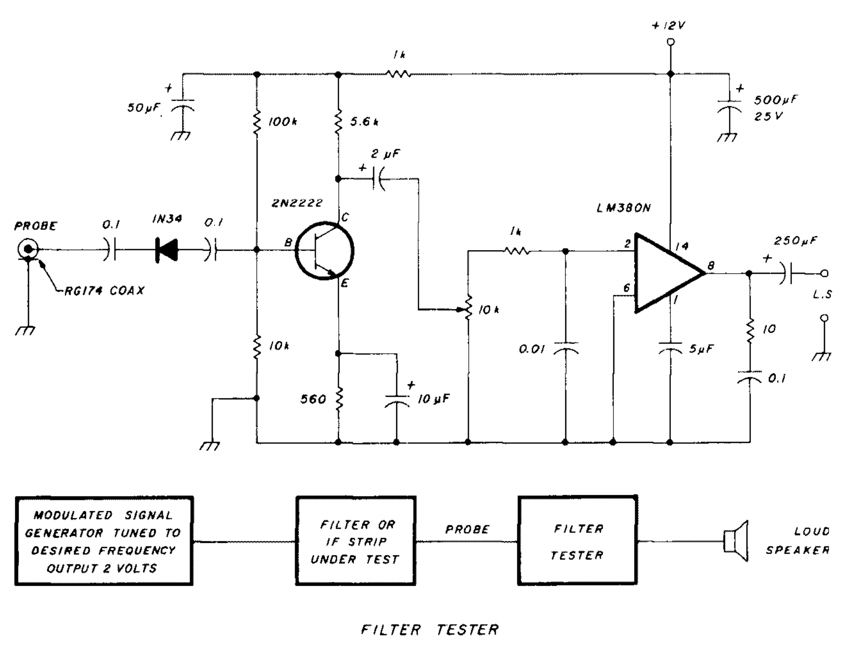
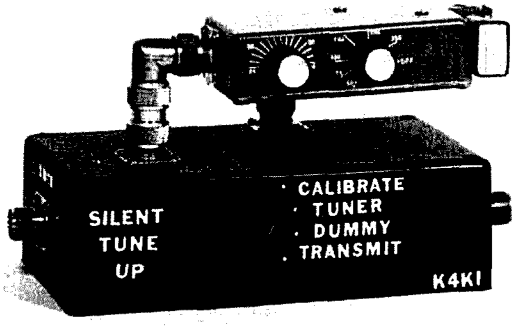


fig. 1. Filter tester.

the
weekender

a safe, silent tuneup



A safe, silent, tuneup — without any radiation of power, and with your SWR at 1:1 at the same time, is readily possible with the unit described in this article.

One of the most useful pieces of equipment is an *accurate* RF bridge; with the advent of small, inexpensive noise bridges, a definite need has been fulfilled. In an effort to obtain maximum utility from such a bridge, the following circuit was developed and built. (The incorporated Palomar noise bridge was loaned to me by my buddy, K4YS, whose comments and advice are a part of this article.)

The noise bridge is mounted "piggy back" on a small chassis box using ordinary coaxial fittings. This mounting provides rigidity, eliminates unnecessary connection cables, and results in a single integrated unit. The circuit is best understood by referring to the schematic of **fig. 1**. S1 (A, B, C, D) is an old CRL, PA-1013, 2-gang, four-circuit rotary switch from my junk box. However, similar switches are available from various manufacturers in sizes appropriate for the power you plan to use. My own Yaesu FT-101-B worked very well with the unit built and described. A defeat circuit section of the switch, S1D, prevents you from accidentally turning on your transmitter when you are in the Calibrate and Tuner position, and possibly damaging your noise bridge.

With the rotary switch, S1B, in the A (Calibrate) position, a 50-ohm 1/2-watt carbon resistor is connected to the UNKNOWN jack of the noise bridge. The RECEIVER jack of the noise bridge is switched through S1A to the *receiver* input. The bridge is balanced by means of the reactance and resistance knobs of the noise bridge for a noise null. Now, when the rotary switch is thrown to the B (Tuner) position, and without touching the already nulled noise bridge control knobs, the antenna tuner is adjusted for a noise null. The tuner is now properly set for a 50-ohm input.

With the rotary switch set to the C (Dummy Load) position, your transmitter can now be properly tuned and loaded for 50 ohms, the nominal value of resistance of your dummy load. And now with the rotary switch thrown in the D (Transmit) position, you are ready to go on the air. As in all RF power switching functions, it is dangerous to engage the rotary switch with the RF power on because of possible transients. However, the transmitter key-up provides an easy way to control transmitter power, and the series defeat circuit of switch S1D provides an additional safety circuit as previously mentioned.

Although a 50 ohm, 1/2-watt carbon resistance is shown for R1, I found that a 47 or 51 ohms value would work as well.

To avoid cutting into the Palomar noise bridge, battery leads to connect to power switch S1C, a couple of battery connectors matching the ones on the noise bridge were connected in order to snap into the circuit as shown. This battery turn-off circuit may at first glance seem to be somewhat redundant, as the resistance knob of the noise bridge has a turn-off switch when the knob is rotated fully counter clockwise. However, if you were to do this, you would unbalance your noise bridge, and if you later switched the bridge back on, the unbalanced condition would make a loud and uncomfortable noise. The bridge could be rebalanced to eliminate this noise, but the power turn-off switch makes operation easier.

My own experience has been that after you're properly loaded up, and your antenna tuner is properly set, your SWR is 1:1, or so close to it that you do not have to do any of the "tweaking" sometimes found necessary with other systems. Safety is quite important; if you keep your transmitter SWR down to 1:1, you'll avoid the arc over problems that sometimes occur when your tune-up SWR gets out of control.

No problems were encountered in the building or testing of the unit. The jacks J1 through J5 were internally connected together with a piece of No. 12 copper antenna wire to provide a common ground at the shells. As a precaution, the leads from switches S1C and S1D were twisted to prevent RF pickup, although none was evident at any time.

I'll be pleased to answer all questions — just send along an SASE to me at 1245 S. Orlando Avenue, Cocoa Beach, Florida 32931.

By William Vissers, K4KI, 1245 S. Orlando Avenue, Cocoa Beach, Florida 32931

item	description
J1-5	SO-239 UHF panel receptacle
R1	50-ohm non-inductive resistor
S1	4-section, 4 position rotary switch
chassis	7-3/4 × 4-1/4 × 2-1/2 inches

Two PL-259 coax adapters, double male; two 646 or 83-1AP coax right angle adapters (used to mount Palomar noise bridge to J4 and J5 as shown in photograph).

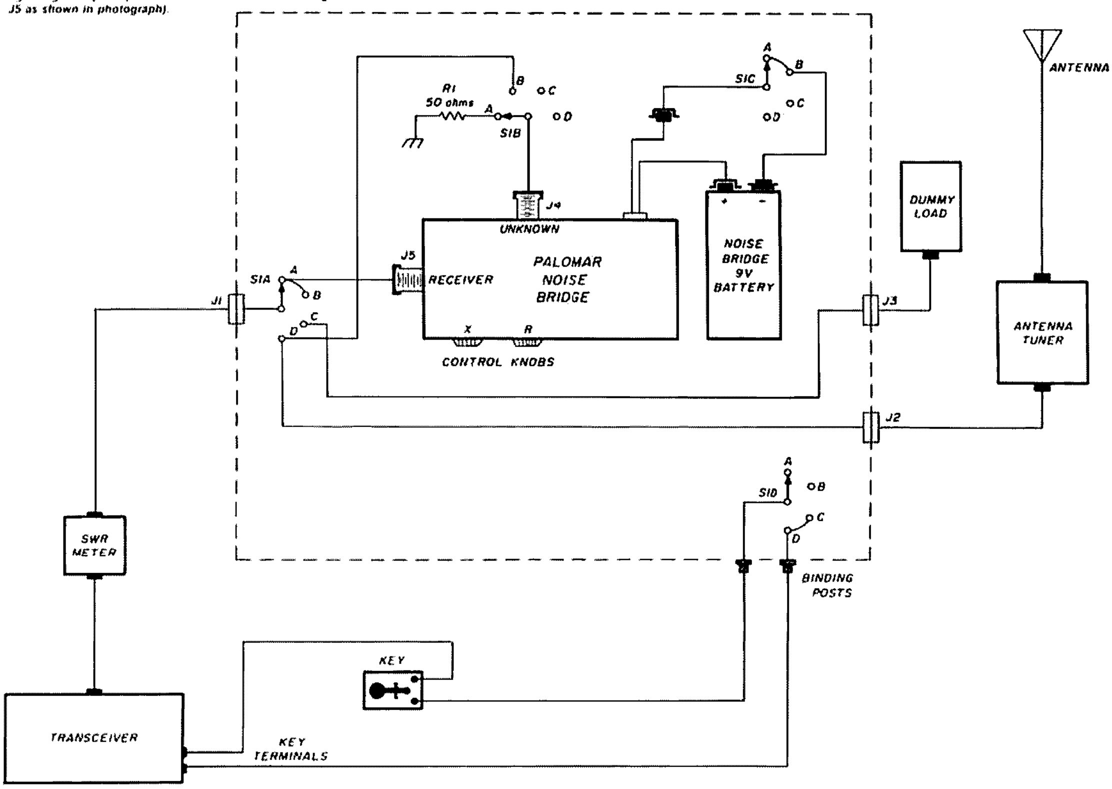


fig. 1. Silent tune-up accessory unit schematic and interconnection diagram.

ham radio

ham radio cumulative index

1980-1984

antennas and transmission lines

general

high-frequency antennas

vhf antennas

matching and tuning

towers and rotators

transmission lines

commercial equipment

construction techniques

digital techniques

features and fiction

fm and repeaters

hazards

integrated circuits

keying and control

measurements and test equipment

microprocessors, computers and calculators

miscellaneous technical

novice reading

operating

oscillators

Short circuit p. 125, May 84
RF synthesizers for hf communications, part 2
WA6OAA p. 48, Sep 83
Short circuit p. 125, May 84
RF synthesizers for hf communications, part 3
WA6OAA p. 17, Oct 83
Short circuit p. 125, May 84
VFOs tuned by cylinder and disc
WØYBF p. 58, Feb 83
Wideband VCO design
WA4MGX p. 49, Jul 84
10 GHz oscillator, ultra stable
K8UR p. 57, Jun 83

power supplies

AC converter, DC to 400-Hz (HN)
WB2YVY p. 58, Mar 83
Battery charging (letter)
Carlson p. 6, Nov 80
Bench power supply (weekender)
WB6AFT p. 50, Feb 80
Diesel generator repair (Tech. forum)
Richardson, Wayne p. 46, Apr 83
Drake R-4C receiver improved power supply
W3RJ p. 28, Feb 82
Dual voltage power supply
WD4SKH p. 32, Mar 83
Comments, WB2UAQ p. 12, Jul 83
Short circuit p. 80, Jul 83
Dual voltage surge-protection for high-voltage power supplies (weekender)
K8VIR p. 42, Aug 81
Electrolytic capacitors (letter)
WB8MKU p. 6, Jun 81
Forget memory (Ni-Cd discussion)
KØOV p. 62, Jan 83
Power supply, amplifier
WA2GFP p. 32, Sep 83
Power supply for the big amplifier
W6YUY p. 64, Jun 82
Power supply, six-output
Martin, Vaughn D. p. 12, Oct 84
Power supply, switching high-voltage
W5FG p. 48, Apr 84
Protection for your solid-state devices
W1OOP p. 52, Mar 81
Regulator problem solved (HN)
W6XM p. 97, Dec 83
Safe power for your low-noise GaAs FET amplifier
WA9HUV p. 18, Nov 82
Squirrel-cage motors make field-day power supplies (HN)
K6DZY p. 74, Aug 81
Trans-global power supply (HN)
W9CGI p. 76, Nov 82
Two-way power for the IC2AT 2-meter handheld
WB3JJF p. 57, Feb 82
Comments WB4MNW, WB3JJF p. 8, Jul 82
Vacuum tube substitution
W2YE p. 58, Oct 83
VHF transceivers, regulated power supply for
WA8RXU p. 58, Sep 80

propagation

Digital Ionosondes
K2RR p. 14, Dec 83
DX forecaster
KØRYW p. 76, Jan 81
DX forecaster
KØRYW p. 92, Feb 81
DX forecaster
KØRYW p. 78, Mar 81
DX forecaster
KØRYW p. 52, Apr 81
DX forecaster
KØRYW p. 76, May 81
DX forecaster
KØRYW p. 52, Jun 81
DX forecaster
KØRYW p. 56, Jul 81
DX forecaster
KØRYW p. 46, Aug 81
DX forecaster
KØRYW p. 48, Sep 81
DX forecaster
KØRYW p. 46, Oct 81
DX forecaster
KØRYW p. 76, Nov 81
DX forecaster
KØRYW p. 78, Dec 81
DX forecaster
KØRYW p. 74, Jan 82
DX forecaster
KØRYW p. 74, Feb 82
DX forecaster
KØRYW p. 82, Mar 82
DX forecaster
KØRYW p. 70, Apr 82
DX forecaster
KØRYW p. 50, May 82
DX forecaster
KØRYW p. 42, Jun 82
DX forecaster
KØRYW p. 78, Jul 82
DX forecaster
KØRYW p. 80, Aug 82
DX forecaster
KØRYW p. 82, Sep 82
DX forecaster
KØRYW p. 82, Oct 82
DX forecaster
KØRYW p. 84, Nov 82
DX forecaster
KØRYW p. 80, Dec 82
DX forecaster
KØRYW p. 74, Jan 83
DX forecaster
KØRYW p. 56, Feb 83
DX forecaster
KØRYW p. 84, Mar 83
DX forecaster
KØRYW p. 94, Apr 83
DX forecaster
KØRYW p. 74, May 83
DX forecaster
KØRYW p. 65, Jun 83
DX forecaster
KØRYW p. 82, Jul 83
DX forecaster
KØRYW p. 66, Aug 83
DX forecaster
KØRYW p. 84, Sep 83
DX forecaster
KØRYW p. 87, Oct 83
DX forecaster
KØRYW p. 90, Nov 83
DX forecaster
KØRYW p. 92, Dec 83
DX forecaster
KØRYW p. 83, Jan 84
DX forecaster
KØRYW p. 79, Feb 84
DX forecaster
KØRYW p. 93, Mar 84
DX forecaster
KØRYW p. 93, Apr 84
DX forecaster
KØRYW p. 119, May 84
DX forecaster
KØRYW p. 109, Jun 84
DX forecaster
KØRYW p. 103, Jul 84
DX forecaster
KØRYW p. 63, Aug 84
DX forecaster
KØRYW p. 79, Sep 84
DX forecaster
KØRYW p. 100, Oct 84
DX forecaster
KØRYW p. 92, Nov 84
DX forecaster
KØRYW p. 63, Dec 84
Dxing by computer
NS6N and Buchanan, Walter p. 81, Aug 84
Grayline propagation, fundamentals of
KR7L p. 77, Aug 84
Moon-tracking by computer
K6WX p. 38, Mar 84
New band propagation
KK2XJM (W4MB) p. 12, Jan 83
Radio signals, radiation of
W1GV/4 p. 26, Jun 82
Russian Woodpecker, the: a continuing nuisance
KR7L p. 37, Nov 84
VHF meteor scatter communications
AI7J p. 69, Feb 84
VHF/UHF world: improving meteor scatter communications
W1JR p. 82, Jun 84
VHF/UHF world: the VHF/UHF Primer, an introduction to propagation
W1JR p. 14, Jul 84
VHF/UHF world
W1JR p. 45, Sep 84

receivers and converters

general

Active mixers, performance capability: part 1
DJ2LR p. 30, Mar 82
Active mixers, performance capability: part 2
DJ2LR p. 38, Apr 82
Audio processor, communications, for reception
W8NRW p. 71, Jan 80
Automatic repeater/receiver sensitivity (HN)
VE7ABK p. 81, Jan 82
Auto-product detection of double-sideband
K4UD p. 58, Mar 80
Letter G3JIP p. 6, Oct 80
Bragg-cell receiver
WB3JZO p. 42, Feb 83
Cascaded stages, IMD and intercept points of
W3IMG p. 28, Nov 84
Communications receiver
K2BLA p. 12, Jul 82
Communications receivers, calculating the cascade intercept point of
WA7TDB p. 50, Aug 80
Crystal ladder filters, systematic design of
N7WD p. 40, Feb 82
CW filter, high performance
W3NQN p. 18, Apr 81
Comments W3NQN p. 6, Nov 81
Detector, logarithmic with post-injection marker generator
W1ERW p. 36, Mar 80
LF converter, fixed-tuned
K1RGO p. 19, Jan 83
Low-noise preamplifiers with good impedance match
W1OOP p. 36, Nov 82
Measuring receiver dynamic range: an addendum (HN)
WB6CTW p. 86, Apr 81
Multiple receivers on one antenna (Two for one) (HN)
W2OZY p. 72, Jun 80
Noise discriminator, a pulsewidth
W6NRW p. 23, Nov 84
Noise figure relationships (HN)
W6WX p. 70, Apr 80
Panoramic adaptor/spectrum analyzer design notes
WA6NCX/1 p. 26, Feb 83
Comments, K2CBY p. 12, Sep 83
Short circuit p. 70, Oct 83
Preamp at work, Quiet!
N6TX p. 14, Nov 84
Receiver dynamic range
W3JZO p. 77, Dec 82
Receiver dynamic range (letter)
AA6PZ p. 7, Aug 80
Remote-site receivers and repeater operation
K9EID p. 36, Jan 83
Rotary dial and encoder for digital tuning
N3CA p. 30, Dec 82
Signal-strength, measuring
W2YE p. 20, Aug 80
Spectrum analyzer, a handheld optical
WA4WDL p. 23, Apr 84
Talking clock (letter)
N9KV p. 75, Feb 80
Talking digital readout (letter)
N5AF p. 6, May 80
VHF/UHF world — high dynamic range receivers
W1JR p. 97, Nov 84

high-frequency receivers

Blanking the Woodpecker: part 1
VK1DN p. 20, Jan 82
Blanking the Woodpecker, part 2: a practical circuit
VK1DN p. 18, Feb 82
Comments, NP4B p. 8, Jul 82
Blanking the Woodpecker, part 3: an audio blanker
VK1DN p. 22, Mar 82
CB to 10 fm transceiver conversion
VE3FIT, VE3AQN p. 16, Feb 83
Communications receivers, high frequency, recent developments in circuits and techniques for
DJ2LR p. 20, Apr 80
Communications receivers for the year 2000, part 1:
DJ2LR p. 12, Nov 81
Communications receivers for the year 2000, part 2:
DJ2LR p. 36, Dec 81
Compact SSB receiver
K1BQT p. 10, Nov 83
Comments KAØGPE p. 10, Mar 84

high-frequency transmitters

vhf and uhf transmitters

vhf and microwave

general

vhf and microwave antennas

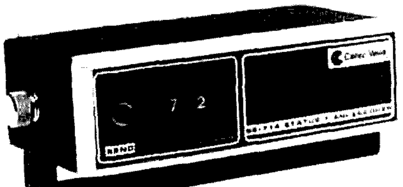
vhf and microwave receivers and converters

vhf and microwave transmitters

NEW products

mobile encoder

The new Model SE-714 mobile encoder from Cetec Vega, when triggered, produces a 3-to-6-digit DTMF ANI (automatic number identification) message plus a 2-digit status message. The ANI message is field-programmable by means



of a diode matrix. The status message is operator-programmable by means of two panel-mounted back-lit thumbwheel switches, allowing up to 100 different status messages.

The SE-714 has two timing features. One feature inhibits the message for a timed period after message transmission, allowing short voice responses without the message. The other ("stuck-mic") feature automatically sends the message after a timed period of continuous transmission.

The SE-714 expands channel capacity by replacing mobile-to-base voice responses with a complete ANI plus status message that requires only a 3/4-second transmission time. The message may be initiated by PTT (or off-hook) operation or by a manual pushbutton on the front panel. The messages can be displayed on any of several types of DTMF displays available from Cetec Vega.

For additional information, contact Cetec Vega, 9900 Baldwin Place, El Monte, California 91731.

Circle #301 on Reader Service Card.

adhesive/sealant and dielectric compound

General Electric Silicone Products Division is introducing RTV-108 Adhesive/Sealant and G-635 Dielectric Compound for the protection of components from the corrosive effects of water, ozone, oxidation, and chemicals and to keep RFI leakage to a minimum.

RTV-108 is a paste-like sealant that cures at room temperature to a flexible silicone rubber. When poured over connectors, taps, and other components, RTX-108 renders them virtually immune to the elements. The product is preferable to heat shrink tubing, especially in underground systems where the equipment is constantly exposed to water.

Because of silicone's unique molecular structure, RTV-108 bonds easily to most clean surfaces without the aid of primers. The product can withstand temperatures from −70 to 400 degrees F without becoming brittle or melting.

G-635 Dielectric Compound, unlike RTV (room temperature vulcanizing) sealants, maintains its soft-to-medium consistency when exposed to the environment. The product is used as a lubricant on aluminum connector threads to prevent seize-up due to oxidation or the bonding of dissimilar metals. When applied as a grease to F fittings, taps, traps, ground blocks, feeder/connector gaps and house splitters, G-635 protects them from the attacks of airborne pollutants such as salt, moisture, and acid rain.

G-635 Dielectric Compound maintains its spreadability at −100 degrees F and shows minimal bleed up to 450 degrees F, making it an ideal lubricant under adverse conditions. A light coat of G-635 on amplifier gaskets keeps them from drying out and splitting, thereby insuring that RFI leakage is kept to a minimum. The product is also used on high-voltage industrial, automotive, and aircraft applications.

For more information about GE silicone rubbers and lubricants, contact Silicone Products Division, General Electric, Waterford, New York 12188.

Circle #302 on Reader Service Card.

legal limit amplifier

The Titan 425 Linear Amplifier from Ten-Tec delivers the full new legal power limit of 1500 watts PEP SSB output and 1500 watts of full break-in power for QSK, CW or AMTOR. This cool-running design consists of two sections, an amplifier and a power supply. Styled to match modern transceivers, and extremely compact for its ratings, the amplifier contains all operating controls and indicators. The power supply is housed in a utility type hide-away enclosure.

The amplifier utilizes two EIMAC 3CX800A7 triodes in a ducted forced air system, operating in a grounded grid configuration easily provide rated output power with up to 65 percent efficiency. Maximum input power of 3 kW (2 kW CCS) requires only 100 watts of drive power. A high-low plate voltage switch assures optimum efficiency at lower power (1 kW output) for tune-up or RTTY and SSTV. Three LED status indicators display Standby, Wait, or Operate mode and a fourth LED alerts you when the input is overdriven. Two panel meters provide full-time indication of plate current and switch-selected choice of plate voltage, grid current, forward power or reflected power. Peak power is indicated on a ten-element LED bar graph display. Band coverage of 1.8 through 23 MHz Amateur bands is standard. Ten-Tec's export model extends coverage to 29.7 MHz.

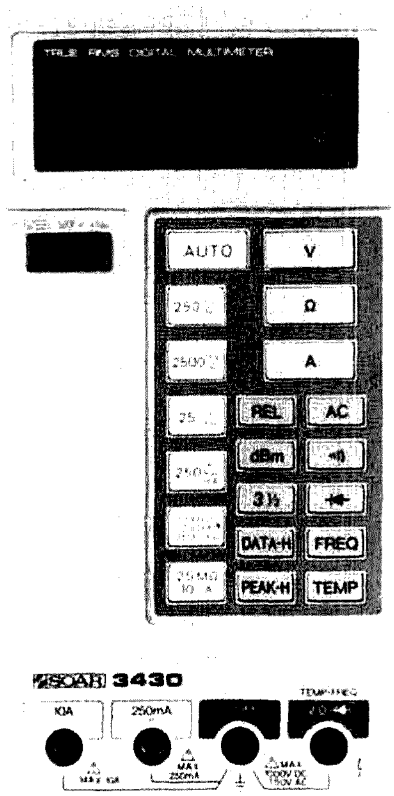
The power supply is conservatively designed for cool operation under full load; the circuit employs a tape-wound Hypersil transformer for minimum weight and size. Primary power of 220-250 volts at 20 amperes is standard. 115 volt operation is possible but not recommended when operating full power. Fuses are provided for primary circuit and plate current. Protective interlocks are included on the AC and high voltage lines.

The Amateur net price of the Titan 425 Linear Amplifier is $2,485.00.

For complete information, contact Ten-Tec, Inc., Sevierville, Tennessee 37862.

multi-function DMM

Designed and developed by North American SOAR Corp., the Model 3430 uses SOAR's new custom LSI chip plus two 4-bit CPU's to achieve 13 different functions and three operational



modes. Unlike the traditional 20,000 count units presently available, the Model 3430 is a 25,000 count instrument. This allows full 4-1/2 digit resolution to the second decimal place for readings such as 240.15 VAC and resistance to 25 megohms. The unit measures True RMS for both AC voltage and current. The unit is priced at $339.

A second DMM, Model 3450, similar to the 3430, is priced at $299.

For details, contact North American SOAR Corp., 1126 Cornell Avenue, Cherry Hill, New Jersey 08002.

Circle #303 on Reader Service Card.

NEW products

climber's belt

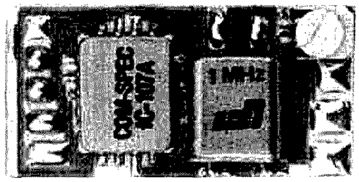
The North Shore Ham Services' Tower Climber's Belt is an all-nylon general purpose safety belt with a 3-inch wide nylon cushion pad and a 1-3/4 inch wide nylon buckle strap. Brass grommet reinforced buckle holes accommodate a single tongue buckle. A 1/2-inch diameter nylon rope lanyard is 3 feet long, making it well suited to use on ham radio towers. The price is $50.00 postpaid.

For additional information, contact North Shore Ham Services, P.O. Box 54, W. Lynn, Massachusetts 01905.

Circle #304 on Reader Service Card.

super-small CTCSS encoder

Communications Specialists of Orange, California, recently announced what it says is now the smallest CTCSS encoder available. The SS-32HB measures only 0.5 × 1.0 × 0.15 inches and will fit into any portable requiring send-only CTCSS. The unit may be programmed to any of the 32 standard sub-audible tone frequencies by bridging solder pads on the board. The price is $29.95.



For further information, contact Communications Specialists, Inc., 426 West Taft Avenue, Orange, California 92665.

Circle #305 on Reader Service Card.

RTTY, SSTV, and Morse programs

A hardware-free RTTY program for Radio Shack™ Model III and IV computers is available from Woodall & Associates. SOFFTY 1.2 provides the most commonly used configuration for RTTY on Amateur bands: 60 and 100 WPM with a 170 Hz tone shift. Keyboard selection of high or low receive tones, as well as normal and inverted code reception are also provided.

SOFTSCAN 1.1 is a no-hardware SSTV decoder and print program for high-resolution hard-copy printouts on the Epson RX series of printers, or on MX's with graphic options.

COMPCODE 1.1B, a hardware-less Morse code program, connects a communications

receiver to a TRS-80™ Model I, III, or IV via the computer cassette interface and cables. Instructions for connecting the receiver to the computer are written into the program itself.

SOFFTY 1.2 sells for $34.95; SOFTSCAN 1.1 for $19.95; and COMPCODE 1.1B for $24.95. Shipping and handling charges are additional.

For complete information, contact Woodall & Associates, P.O. Box 284, Plainfield, Indiana 46168.

Circle #306 on Reader Service Card.

SWL catalog

A new SWL catalog has recently been released by Spectronics describing its complete line of name-brand receivers, antennas, accessories, and publications for the short-wave listener. Written in a non-technical, conversational manner, it is an ideal tool to introduce the beginner to short wave, yet diverse enough to appeal to the more serious listener.

For a free copy, contact Spectronics, Inc., 1009 Garfield Street, Oak Park, Illinois 60304-1890.

dielectric resonator oscillators

Microwave Research & Manufacturing has announced its new line of dielectric resonator oscillators covering 2 GHz to 20 GHz, with standard output powers of up to +23 dBm. Stability is better than 3 PPM per degree over a temperature range of −54 to +100 degrees Celsius. Phase locking ability is optional.

Phase noise is typically −93 dBc at 10 kHz and frequency pulling is less than 0.3 MHz into an output VSWR of 1.5:1. The mechanical tuning range is 8 percent for most models with electrical tuning optional. Power requirements are 5 to 28 VDC and offer high efficiencies of typically 20 percent. Harmonic content is −40 dBc. The size of model MW-7 is extremely small, measuring only 1.42 x 0.76 x 0.600 inches.

Millimeter wave sources are also available using F.E.T./D.S.O. doubling techniques yielding highly stable millimeter wave sources of high efficiencies and small size.

Microwave Research & Manufacturing is a supplier of microwave and millimeter wave FET and Gunn diode dielectrically stabilized oscillators, VCOs and downconverters.

For additional information, contact D.C. Mitchell, Microwave Research & Manufacturing Company, 1 Cider Mill Lane, Upton, MA 01568.

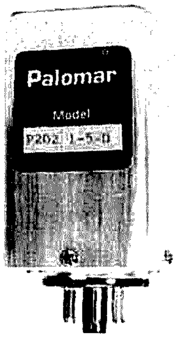
Circle #307 on Reader Service Card.

two-digit DTMF decoder

A new two-digit sequential DTMF decoder is available from Palomar Engineers. The decoder, Model P-202, features 0.5 ampere 115 volt relay contacts, dual bandpass filters, quartz crystal frequency control, as well as operation over an ex-

NEW products

tended temperature range and over a wide range of signal levels. It is highly immune to false triggering. Operation is from 12 to 30 volts DC. DTMF coding is factory-set to order but easily changed by moving circuit board jumpers. The unit is priced at $125.

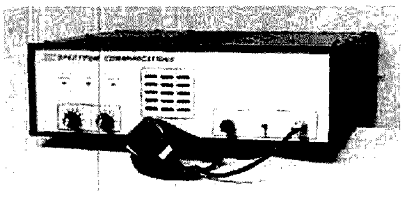


For more information, contact Palomar Engineers, 1924-F West Mission Road, Escondido, California 92025.

Circle #308 on Reader Service Card.

desktop/portable repeater

The SCR77D from Spectrum Communications is a new "Decktop/Portable" Repeater. The first of its kind on the *Amateur* market, the 77D was developed in response to customer demand over the past several years. Its compact, low-power configuration makes it suitable for "local" use (within a 0-20 mile radius, depending on antenna and terrain). It may be used at a fixed location, or Portable/Mobile.



Powered by a 12 volt battery pack, the SCR77D can be mounted in a vehicle parked on a hilltop for temporary coverage as needed. Autopatch and 'PL' are included. An AC power supply, as well as jacks for 12 VDC power, are built in. Full duplex base station applications, such as computer data links or export "rural telephone," are also ideal for the SCR77D. Standard models include 10 watt UHF unit with built-in duplexer and a 15-watt VHF unit with external duplexer.

For details, contact Spectrum Communications Corporation, 1055 W. Germantown Parkway, Norristown, Pennsylvania 19401-9616.

Circle #309 on Reader Service Card.

flea market

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HAMFESTS Sponsored by non-profit organizations receive one free Flea Market ad (subject to our editing) on a space available basis only. Repeat insertions of hamfest ads pay the non-commercial rate.

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IBM-PC RTTY ASCII/BAUDOT/CW send and receive. Split screen, buffers and features beaucoup. SASE to: E. Alline, 773 Rosa, Metairie, LA 70005.

CX7 REPAIRS. 415-549-9210.

HW-101 XCVR, built-in digital display. RIT, CW filter, HP-23A supply, HP-13A mobile supply, $350. BC-224-D rcvr, w/spply $50. Heathkit IG-42 sig-gen $25. Vibroplex "champion" $25. (509) 928-3418, WD6DXX.

TRAVEL-PAK QSL KIT — Converts post cards, photos to QSLs. Stamp brings circular. Samco, Box 203-c, Wynantskill, New York 12198.

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Republic of South Africa

WANT old antenna books, handbooks, CQ, pre-1940 QST. Tad Cook, KT7H, 5519 — 12th N.E., Seattle, WA 98105.

CUSTOM MADE embroidered patches. Any size, shape, colors. Five patch minimum. Free sample, prices and ordering information. Hein Specialties, Inc., Dept. 301, 4202 N. Drake, Chicago, IL 60618.

OLD antenna or radio books wanted, pre-1940, the older the better. Clem, KR6A, 26530 Parkside Drive, Hayward, CA 94542. (415) 886-1205.

WANTED: Defective or operating National LF-10 low frequency converter. Harry Weber, 2605 West 82nd Place, Chicago, IL 60652.

RUBBER STAMPS: 3 lines $4.50 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

COLLINS KWM-380 with all optional interfaces. Perfect $2495. WA6NWP, Sid, 209-642-3363. 52632 RD 426, Oakhurst, CA 93644.

SAMCO introductory offer. Roll of 100 b/w call strips. $1.50 pp. SAMCO, PO Box 203, Wynantskill, NY 12198.

KEYER/CODE trainer chip. One evening project. Great project for beginners, clubs or anyone who needs a good, low cost iambic keyer or code trainer. $15 ea. PP. VISA/Mastercard accepted. Micro Digital Technology, PO Box 1139, Mesa, AZ 85201. (602) 897-2534.

REPAIR, ALIGNMENT, calibration. Collins written estimates $25; non Collins $50. K1MAN. (207) 495-2215.

ATLAS 350XL owners group. Send QSL card with s/n your rig. Know anyone who repairs them? Have any technical information to share? Any questions? Rod, N5NM, Box 2169, Santa Fe, NM 87504.

CHASSIS and cabinet kits. SASE K3IWK.

SCHEMATICS: Radio receivers 1920/60's. Send name brand, model, SASE, Scaramella, P.O. Box 1, Woonsocket, RI 02895-0001. (602) 897-2534.

AMATEUR RADIO's newspaper — Worldradio. Latest info. One year subscription (12 issues) only $10. Worldradio, 2120-B 28th St., Sacramento, CA 95818.

DIGITAL automatic displays for FT-101's, TS-520's (functional DG-5 replacement), Collins, Drake, Swan, Heath and most others. Six 1/2" digits in a 5" by 1-1/2" × 9" cabinet! Write for information. Grand Systems, PO Box 2171, Blaine, Washington 98230. (604) 530-4551.

ENJOY SATELLITE TELEVISION. Save money with easy, guaranteed, do-it-yourself antenna plans/kits. Electronic knowledge not necessary. Send $1.00 for catalog or $8.95 for "Consumer Guide to Satellite Television". GFI-41, Box 9108, Missoula, MT 59807.

LARSEN ANTENNAS at low prices. Most models in stock. Shipped UPS. NCN Electronics, 33 McKinley Avenue, West Orange, NJ 07052. (201) 731-9506.

ELECTRON TUBES: Receiving, transmitting, microwave . . . all types available. Large stock. Next day delivery most cases. Daily Electronics, 14126 Willow Lane, Westminster, CA 92683. (714) 894-1368.

CUSTOM embroidered emblems — Cloisonne enameled pins, your design, low minimum, excellent quality, free booklet. A.T. Patch Co., Dept. 65, Littleton, NH 03561. (603) 444-3423.

FOR SALE: GR-1606A RF impedance bridge. For serious MF/HF antenna work, $700. Serious inquiries/offers please! John Lenahan, KØRS, 1202 Jay St., Ottumwa, Iowa 52501. (515) 682-0711.

WANTED: Cash paid for used speed radar equipment. Write or call: Brian R. Esterman, PO Box 8141, Northfield, Illinois 60093. (312) 251-8901.

$$$$ SUPER SAVINGS on electronics parts, components, supplies, and computer accessories. Free 40-page catalog for SASE. Get on our mailing list. BCD Electro, PO Box 830119, Richardson, TX 75083. Or call (214) 690-1102.

WANTED: Old microphones, remote mixers other misc related items. All pre-1935. Box Paquette, 107 E. National Avenue, Milwaukee, WI 53204.

WANTED: Used microwave equipment. D. Mitchell, K8URL, 1 Cider Mill Lane, Upton, MA 01568. (617) 529-4638.

RECONDITIONED TEST EQUIPMENT $1.00 for catalog. Walter, 2697 Nickel, San Pablo, CA 94806.

RTTY-EXCLUSIVELY for the Amateur Teleprinter. One year $7.00. Beginners RTTY Handbook $8.00 includes journal index. P.O. Box RY, Cardiff, CA 92007.

"HAMS FOR CHRIST." Reach other Hams with a gospel tract sure to please. Clyde Stanfield, WA6HEG, 1570 N. Albright, Upland, CA 91786.

TENNATEST — Antenna noise bridge — out-performs others, accurate, costs less, satisfaction guaranteed, $41.00. Send stamp for details, W8URR, 1025 Wildwood Road, Quincy, MI 49082.

WANTED: Early Hallicrafter "Skyriders" and "Super Skyriders" with silver panels, also "Skyrider Commercial", early transmitters such as HT-1, HT-2, HT-8, and other Hallicrafter gear, parts, accessories, manuals. Chuck Dachis, WD5EOG, The Hallicrafter Collector, 4500 Russell Drive, Austin, Texas 78745.

VERY in-ter-est-ing! Next 4 issues $2. Ham Trader "Yellow Sheets", POB356, Wheaton, IL 60189.

Coming Events

ACTIVITIES

"Places to go..."

WISCONSIN: The "original" annual midwinter Swapfest, Saturday, January 12, 8 AM, Waukesha Co. Expo Center Forum, Milwaukee. Admission $2.00 advance, $3.00 door. Reserved 4' tables $3.00 prior to December 31, $4.00 at door. Sponsored by the West Allis Radio Amateur Club. For tickets: WARAC Swapfest, PO Box 1072, Milwaukee, WI 53201. SASE please.

MINNESOTA: The annual Handi-Ham Winter Hamfest, Saturday, December 1, Eagles Club in Faribault. Registration 9 AM. Handi-Ham equipment auction. Dinner at noon. Talk in on 19/79. For information: Don Franz, WØFIT, 1114 Frank Avenue, Albert Lea, MN 56007.

INDIANA: South Bend Hamfest Swap & Shop, January 6, first Sunday after New Year's Day at Century Center downtown on US 33 Oneway North between St. Joseph Bank Building and river. Industrial history Museum in same building. Carpeted half acre room. Open tables $1 per ft. Four lane highways to door from all directions. Talk-in freq: 52-52, 99-39, 93-33, 78½18, 69-09, 145.29.

VIRGINIA: The 8th annual Richmond Frostfest, sponsored by the Richmond Amateur Telecommunications Society. Sunday, January 13, Virginia State Fairgrounds. 8:30 AM to 3:30 PM. General admission $4.00. Flea market spaces $3.00 without table; $7.00 with 8' table. Entire show indoors. Deadline for booth, December 30; flea market, January 10. Building open Saturday afternoon for setup. Write Richmond Frostfest, PO Box 1070, Richmond, VA 23208 or call Bill Scruggs, N4DDM (804) 272-8206.

OPERATING EVENTS

"Things to do..."

THE SANDY RIVER ARC will operate the Chester N. Greenwood special event station in commemoration of the inventor of the Earmuff. 1500Z December 21 to 2100Z December 23. 10K up from the General split on 80, 40 and 20M. QSL with 40¢ postage to KA1CNG via Callbook for special 8 x 10 certificate.

A SPECIAL EVENT STATION commemorating the 63rd anniversary of the Tuscaloosa Jaycees on Saturday, January 12. KE4TN will operate from 13 to 23 Zulu and will offer an 8½ x 11 certificate to all contacts. Send QSL card to the Tuscaloosa Jaycees, PO Drawer L, Tuscaloosa, AL 35404 or Callbook address of KE4TN.

MERSEYSIDE SPECIAL EVENT GROUP will be operating during the month of December special event call sign block GBO 1, 2, 4, 6 and 8 BCL. (Beatle City Liverpool) from 00.00 December 1 to 24.00 December 31. The group will operate all HF bands and 2m and 70cm, all modes. The special event calls are to celebrate the opening of the Beatle City Museum in Liverpool. QSL information direct only to: QSL Manager, G4VKV, c/o Beatle City, PO Box 12, Liverpool, England. Special QSL cards depicting the Beatle City Museum and special event status will be available for all verifiable QSO's and all listener reports.

THE GUERRI REPORT

Ernie Guerri
W6MGI

digital signal processing is just around the corner

Applications and techniques for signal processing have become complicated and diverse. An increasing number and variety of integrated circuits whose specialized functions are related to signal processing tasks are now available.

The basic task of signal processing is enhancement of signal-to-noise ratio. The "noise" can be electronic, acoustic, optical, even mathematical. This last aspect has led to modern digital signal processing techniques.

Consider, for example, a television picture: in its analog form, a single frame will have an almost infinite number of shades of gray — every shade from white to black. Some portions of the picture signal may include undesired information such as ignition noise, snow, ghosts, or sync instability. If we were to sample this picture in some manner that would convert the analog signal into a numerical (digital) signal, it would then be possible to assign the undesired information a specific numeric value and address. If we had a way of knowing which pieces of information were intended to be part of the actual picture, then all other pieces could be assumed to be extraneous, and removed. This is the basic technique used in assembling pictures from information sent back from planetary probes, in removing noise and ghosts from network television transmissions, and in compressing the bandwidth needed for transmitting real-time video.

Recently I saw a demonstration in Japan of a TV ghost eliminator made by Toshiba. Multiple images created by tall buildings in an urban environment were completely eliminated using advanced signal processing techniques in the TV set. At the moment, the cost is high — but the technology is here. It won't be too much longer before we have the capability for transmitting acceptable quality real-time video in the space needed for only a few 2-meter FM signals.

higher frequencies

In examining frequency band utilization in decade steps (220 MHz, 2300 MHz, 25 GHz, and 250 GHz) it appears that the higher one goes, the less experimental activity one finds. Yet it's at 250 GHz that the state-of-the-art is actively being explored. The results of this activity are surprising. It takes every trick in the book to get a front-end working at these frequencies, but a team at the National Radio Astronomy Observatory has fabricated a fundamental frequency mixer with a noise figure of less than 9 dB. This device operates at 21 degrees K, using a Schottky diode with a connecting whisker 90 μm (1 μm = 10^{-6} meter) long. The RF bandwidth is over 70 GHz, and IF bandwidth can be up to 10 GHz. Each of these devices can therefore handle an information bandwidth equal to the entire commercial electromagnetic spectrum we now use!

At present, no corresponding devices exist for the generation and amplification of RF energy which would make regular operation at these frequencies possible. But the incentive is there: the spectrum around 260 GHz is a "window" in the atmosphere in which microwave attenuation is only about 5 dB/km.

"faceless" test equipment

We're accustomed to test equipment dedicated to specific measurement tasks. We bring an ailing unit *to* a location — usually a test bench — loaded with meters, generators, analyzers, and such. However, in many major technical companies this traditional approach is being replaced by distributed instrumentation, using microcomputers for display and data processing, with several different "front ends" at hand to accomplish the desired test. One company now offers a dual-channel 50-MHz digital storage "scope" for the Apple computer, and a major instrument supplier has announced a 128-channel waveform digitizer, featuring 0.1 percent accuracy, which multiplexes its data channels to an IBM PC.

This approach should be useful for modular instruments that replace those based on the plug-in concept. Users of very complex instruments such as spectrum analyzers will benefit from the processing power available in this new approach. Although cost will be a limitation at first, we can expect that portable and field test equipment will eventually make use of this approach. Think of the benefit to the service technician with a whole library of diagnostic data available in a single multifunction scopevoltalyzer!

Amateurs have long lacked access to accurate measurements in the frequency domain. The advent of affordable digital signal processing could put a quality spectrum analyzer into the typical Amateur installation within the forseeable future.

ham radio